JULY · 1954

roceedings of the I·R·E

Journal of Communications and Electronic Engineering

IRE WESTERN CONVENTION

Volume 42 Number 7

n Pacific Auditorium, shown above, and the Ambassador Hotel geles will be the scene of the Western Electronic Show and Con- WESCON) on August 25-27. Sponsored annually by the IRE and it is one of the largest engineering conventions of the year. (See 4 for details.)

IN THIS ISSUE

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IRE Standards on Television Measurements and on Audio Terms appear in this issue.

The Institute of Radio Engineers


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COMPACT
Ouncer
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Magnetic Amplifiers
Multi-Shielded Inputs
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HERMETIC... MIL-T-27
Audios
Pulse Units
TO SPECIFICATIONS...
UNITED TRANSFORMER COMPANY • 150 VARICK STREET, NEW YORK 16, N. Y.

NEWS and NEW PRODUCTS

July 1954

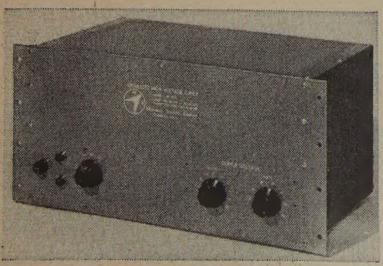
Clinical Defibrillator

Instantly operative when turned on, the Morris Clinical Defibrillator manufactured by **Levinthal Electronic Products, Inc.** provides flexible facilities for direct application of electric shock to the ventricles of the heart in surgical cases involving ventricular fibrillation. The output of several amperes of 60-cps alternating current is applied to stainless steel electrodes which are completely insulated except for their active faces. Circuitry includes isolation transformer and other safety features.

Three automatically-controlled pulse lengths of 0.1, 0.5, and 1.0 seconds are augmented by manual facilities where the pulse length is directly under control of the surgeon. In one model, five steps of pulse amplitude are delivered: 60, 85, 110, 135, and 160 volts. Another model provides continuously-variable amplitude from 0 to 160 volts. Shock rate, either single or serial, is controlled by the surgeon either from the panel switch or from an accessory foot switch specially sealed for use in the presence of inflammable or explosive anesthetic gases.

High-Voltage Power Supply

Northeast Scientific Co., 1 Gray St., Cambridge, Mass., announces the design and production of a moderately priced, wide-range regulated power supply for gen-



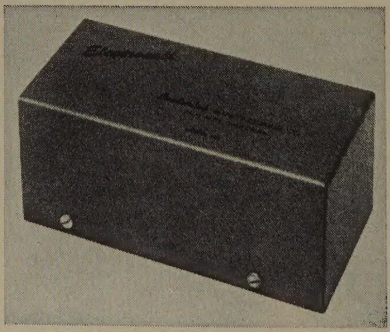
eral laboratory use where a high-voltage source is needed. The Model RE-2002 is

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

suited to applications requiring a highly stable well-regulated, easily adjustable source of high voltage with very low ripple. It can be used with klystrons, which require extreme stability and high current capacity. This power supply has been carefully engineered to be practically immune to line voltage changes and load effects. The use of a balanced, two-stage, high gain feedback amplifier yields a regulator with a stability ratio of 10^6, an internal impedance under 0.5 ohm at dc and a ripple under 0.5 mv, peak to peak. The output voltage is adjustable to ± 2 per cent over the range of 700 to 2,020 volts. Either the positive or negative side may be grounded. The output current rating is 0 to 25 ma. Regulation is maintained at full load for line voltages down to 105 volts and up to 125 volts. It is designed to operate down to 50 cps. All controls are on the front panel. The RE-2002 uses only standard tubes. Price is $350.00.

Fleet-Radio Control Unit

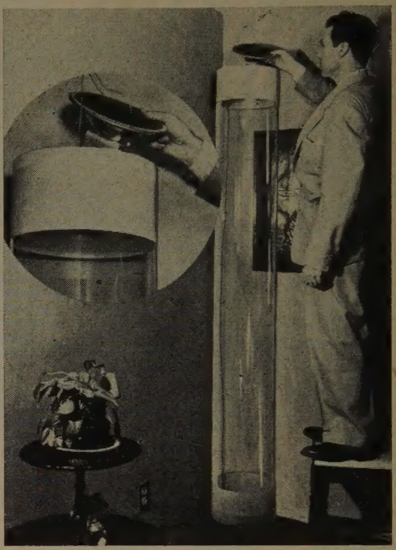
Fleet radio systems may be put on a direct call basis by the addition of a control unit to each of the receivers in the fleet, and installation of a fully automatic coding unit in the base-station transmitter. Manufactured by **Lectrolab Manufacturing Co.**, 2990-C Middlefield Rd., Palo Alto, Calif., the unit is contained in a housing, 2½ by 3 by 7 inches.



In operation, the electrocall unit energizes the output circuits of the receivers upon receipt of the coded signal from the fleet transmitter, thus giving operators of fleet vehicles the advantage of not being required to listen to all transmissions in the frequency band. An accompanying operational advantage is gained through the reduction in battery drain per vehicle of approximately 14 per cent. Price per unit is $49.50.

Sound Box for Speakers

"MorSound," a new plastic housing for 6-inch or 8-inch loud speakers claimed to be a new and improved method of amplifying and clarifying voice and music tone quality is being produced by **Taylor & Art Inc.**, 1710 E. 12th St., Oakland, Calif.



The housing is a cylinder of transparent Kodapak acetate 72 inches long. The speaker is suspended inside one end by mounting to a disc ring. A similar disc ring at the opposite end holds the cylinder in shape. The air column within the cylinder becomes activated by the sound vibrations from the speaker. Low notes are materially reinforced as they traverse the length of the tube before coming out at the lower end. High notes coming from the front of the speaker, positioned within the cylinder to point toward the ceiling, are likewise accentuated. Both the front and the back of the speaker cone are utilized to maximum efficiency because there is no cabinet contamination or cabinet effect in the sound spectrum of the signal induced by the speaker.

As the high notes from one side of the speaker radiate upward to the ceiling and low notes from the back of the speaker roll out the bottom, the distribution of sound in varied directions gives the total tone a new depth and clarity and eliminates the disadvantages of compressing all the sound into a single directional track.

Kits contain all parts with clear and simple directions for assembly. Acetate is rolled into compact package.

(Continued on page 66A)

DIRECT READING
Frequency Meters

***DESIGNED** for easier, more accurate readings...*

Tried and true—the FXR Type 410A Direct Reading Frequency Meters give you bull's-eye accuracy, right to the exact reading spot. No more jiggling of many dials, no more complicated calculations. Human error factor is eliminated. These dependable FXR units provide the most convenient means available for determining oscillator frequency in the microwave region.

- .08% ACCURACY
- FULL WAVEGUIDE FREQUENCY RANGE
- DIRECT READING
- HIGH Q
- NOMINAL 30% ABSORPTION DIP
- BACKLASH-FREE OPERATION

TYPE	FREQUENCY RANGE
C410A	5.85 to 8.20 KMC/S
W410A	7.05 to 10.00 KMC/S
X410A	8.20 to 12.40 KMC/S

DELIVERY FROM STOCK • PRICE $125 EACH*

Write today for new catalog of complete line of FXR Precision Microwave Test Equipment

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VSWR AMPLIFIER
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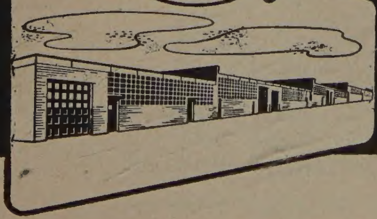


Electronics & X-Ray Division

F-R MACHINE WORKS, Inc.

26-12 BOROUGH PLACE, WOODSIDE 77, N. Y.

*Data and prices subject to change without notice.

IRE People

Donald L. Herr (SM'46) and **Leon H. Weiss** (M'52) have been appointed President and General Manager, and Director of Engineering, respectively, of the recently organized Mark Instrument Company, Inc., in Santa Monica, Calif.

Mr. Herr received the B.S. degree in 1937, the M.S. degree in 1938, and the E.E. degree in 1945, from the Moore School of Electrical Engineering of the University of Pennsylvania. During World War II, he served as a Lt. Commander in the U. S. Navy, during which time he received two letters of commendation for his service. Since the war, he has been affiliated with such companies as the Control Instrument Co. as Assistant Vice President of Engineering, the Reeves Instrument Corp. as Senior Engineer, the Guided Missile Laboratories of the Hughes Aircraft Co. as Technical Advisor, and the American Electronics Manufacturing, Inc. as President. He has also been a consulting engineer in Santa Monica, Calif., and served for one year as an Adjunct Professor of the Graduate School of Brooklyn Polytechnic Institute.

In 1936, Mr. Herr was the author of the AIEE Student Prize Paper "Differential Analyzer." He has also received the A. Atwater Kent Award in 1937, the American Society of Naval Engineers Award in 1945, and the Cressy Morrison Award from the New York Academy of Science in 1950.

Mr. Herr is the author of numerous technical papers and holds several patents in his field. He is a member of the honorary fraternities Sigma Tau and Tau Beta Pi.

Mr. Weiss attended the University of Missouri and the Massachusetts Institute of Technology, and received his B.S. degree from the latter institution in 1942. From 1941–1943 he was a research assistant at M.I.T. During World II he served in the Army Air Force, first as a Technical Assistant in Special Weapons, and later as Liaison to the Jet Propulsion Laboratories at the California Institute of Technology.

After the war he was affiliated with the Guided Missile Laboratories of Hughes Aircraft Co. as a research engineer, and from 1952–1953 he served as Chief Engineer and Chief Sales Engineer for American Electronic Manufacturing, Inc.

He is a member of Eta Kappa Nu.

Homer W. Parker (A'47–M'50–SM'53) has been appointed Manager of the International Testing Service Division of the Jackson and Church Company, Saginaw, Mich. The activities of the organization are devoted to research, development, field and laboratory testing.

H. W. Parker

Mr. Parker received his B.S. degree in Physics in 1947 from North Texas State College, and has sub-

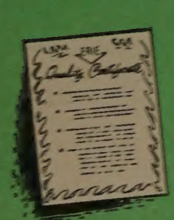
(Continued on page 48A)

An Insurance Policy that Saves You Manufacturing Costs
Included in every shipment of Erie Capacitors

ERIE®

Quality Certificate

What is an ERIE Quality Certificate?



An Erie Quality Certificate is a form that lists the results of both electrical and mechanical tests for every shipment of Erie Capacitors. These tests are made by competent quality control inspectors using modern and precise measuring equipment.

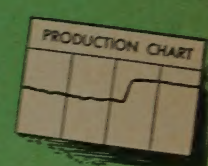
Will it Cut Costs?

YES — With the Quality Certificate you cut costs by reducing incoming inspection. You save the bother, time, and expense of returning faulty material because you are dealing with capacitors of a known quality. You also reduce the risk of putting faulty capacitors in your products.

Here's an Extra Dividend!

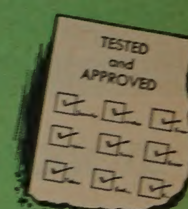
Erie Quality Certified capacitors cost you no more than other kinds. You benefit because quality products are always cheaper to use and add quality to your finished products.

Will it Speed Production?



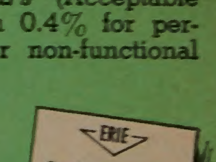
YES — It takes less time for Quality Certified Erie capacitors to get from your receiving doors to your production lines. It eliminates costly trouble-shooting delays on your assembly line and in your inspection of the finished products.

What Does the Quality Certificate Offer?



The Quality Certificate lists the sample size and test results for each inspection sequence or series of inspection tests. The frequency distribution of capacitance values in the sample is also shown. Electrical tests include dielectric strength, insulation resistance, and dissipation factor. Other tests such as temperature coefficient, case insulation breakdown are performed and results listed where applicable. The certificate also contains a complete inspection check list for mechanical and visual items. The sampling tables used are MILITARY STANDARD 105 with AQL's (Acceptable Quality Level) ranging from 0.4% for performance items to 1.5% for non-functional deviations.

Again the Pioneer



As in so many other important developments in electronic components, Erie again leads the field. Erie is the first ceramic capacitor manufacturer to give customers this complete quality information with each shipment.

ERIE TRADE MARK RESISTOR CORP.

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Factories: ERIE, PA. • LONDON, ENGLAND • TRENTON, ONTARIO • HOLLY SPRINGS, MISS.

IRE People

(Continued from page 44A)

sequently pursued advanced studies in physics and electrical engineering at several major universities. During the past 14 years he has devoted a considerable portion of his efforts to defense activities, primarily for the U. S. Navy, including the U. S. Naval Air Missile Test Center, U. S. Naval Ordnance Test Station, and United Aircraft Corporation.

He is a registered professional engineer, a member of the American Physical Society, and the American Institute of Electrical Engineers.

The appointment of **Dr. Robert D. Huntoon** (A'40–SM'47) as Associate Director for Physics, National Bureau of Standards, was recently announced.

Dr. Huntoon was born in Waterloo, Iowa, in 1909 and received his B.A. degree from Iowa State Teachers College in 1932. In 1933 he was awarded a graduate assistantship in the Physics Department of the State University of Iowa, where he specialized in nuclear physics, receiving his Ph.D. in 1938. From 1938 to 1940 he was a member of the New York University faculty, and during the years 1940 and 1941 was a research physicist in the vacuum tube division of Sylvania Electric Products Corporation.

(Continued on page 50A)

IRE People

(Continued from page 48A)

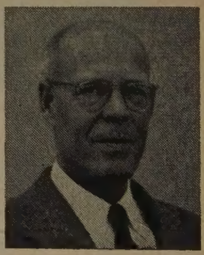
Dr. Huntoon, formerly Director of the NBS Corona (California) Laboratories, brings to the newly-created position of Associate Director of Physics broad experience in the fields of atomic and nuclear physics and electronics. He has also been designated Acting Chief of the NBS Electronics Division and of Bureau's Central Radio Propagation Laboratory.

Dr. Huntoon joined the staff of the NBS in 1941, and was one of their principal scientists concerned with the design and development of the proximity fuze. In 1944 he was loaned to the War Department, where he served as consultant on proximity fuzes and related problems in the Office of the Secretary of War. In this capacity he was sent to the European Theater of Operations as a member of the Advisory Specialists Group, United States Strategic Air Forces.

He was appointed Chief of the NBS Electronics Section in 1945 and directed fundamental research on electronic circuits, control devices, and other electronic ordnance components. In 1947 he became Assistant Chief of the Atomic and Radiation Physics Division, and Chief in 1948. During this period, he also served as Co-ordinator of Atomic Energy Commission projects at the Bureau of Standards.

Dr. Huntoon is a member of Sigma Xi, the American Physical Society, and the Philosophical Society of Washington. In 1948, he received the Washington Academy of Sciences Achievement Award in the Physical Sciences for his contributions to the field of electronics.

The appointment of **Dr. Russell R. Law** (A'34–M'40–SM'43–F'53) to the post of director of research and development at CBS-Hytron has been recently announced. He has been serving as technical advisor on research engineering.



R. R. Law

Dr. Law was born on January 11, 1907, in Hampton, Ia. He is a graduate of Iowa State College and of the Harvard Engineering School. He joined CBS-Hytron in January, 1953, after 18 years with RCA in tube research and development.

Dr. Law is an authority in the field of electron optics and holds many patents in this country and abroad. He is credited with the invention of the darkened glass used almost universally today for the face of modern television picture tubes to improve contrast. Among his applied research achievements have been improvements not only in electron optics, but in the field of transistors, projection screens, pulse triodes, tri-colored picture tubes and UHF television transmitters.

Dr. Law is a member of the American Physical Society.

(Continued on page 52A)

MOMENT

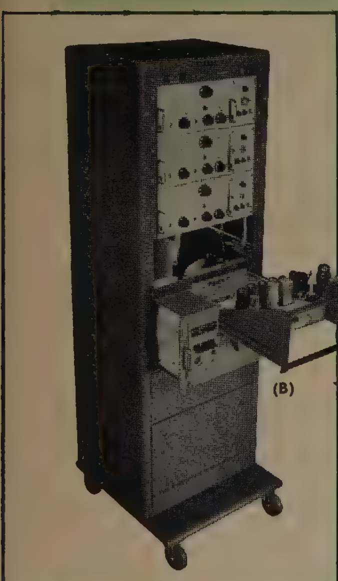
Sanborn "150" records the effects of water wave forces to aid in pile structure design

FORCE

UPSTREAM WAVE GAGE

DOWNSTREAM WAVE GAGE

By means of a Sanborn 150 Oscillographic Recording System equipped with four carrier type preamplifiers, engineers at the M.I.T. Hydrodynamics Laboratory are getting accurate pictures of simulated shallow water waves and their effect on dummy piles. The shape and length of precisely controlled waves in a 90 foot glass flume are plotted simultaneously with their moment and force on a suspended cylindrical pile. The excellent frequency response available with this method permits a sensitivity and accuracy not obtainable in previous model studies of this type.



Preamplifier (A) in position to plug into a driver amplifier with frame, combined with power supply and control panel (B) which is normally already in place in the basic cabinet assembly.

This is but one of MANY applications possible with Sanborn 150 Oscillographic Recording Systems

Virtually all electrical phenomena, within a frequency range of zero to 100 cps, can be accurately, permanently and graphically registered by Sanborn Oscillographic Recording Systems. This versatility of application is possible because of the *flexibility* of Sanborn 150 Series Recording Systems. A wide variety of quickly *interchangeable* preamplifiers, which plug in to *built-in* driver amplifiers (illustrated at left), are available for use with Series 150 Systems, to record such phenomena as: stress, strain, pressure, displacement, thickness, velocity, acceleration, current, voltage, temperature, torque, light, flow, force, load, position, rpm, radiation, tension, and power.

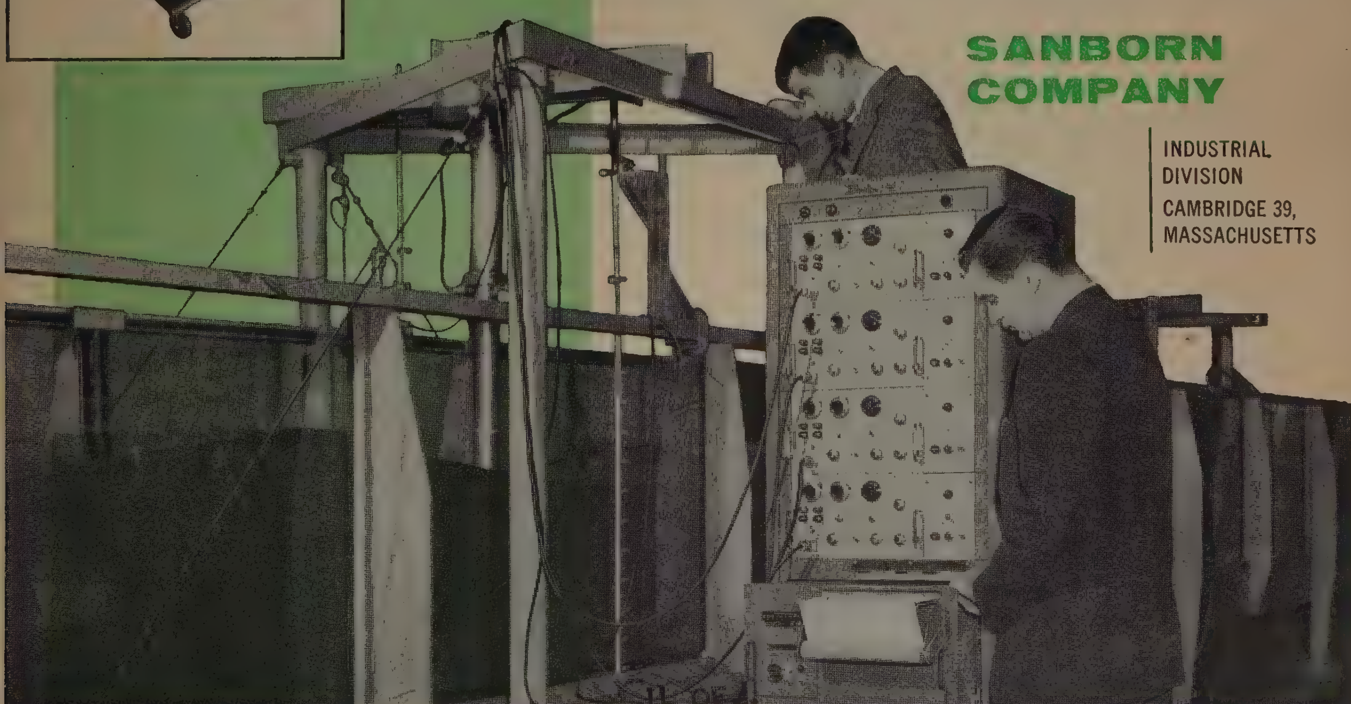
Add to this versatility the Sanborn features of inkless tracings in true rectangular coordinates, on plastic coated chart paper . . . high torque movement . . . time and code markers . . . numerous chart speeds.

Let Sanborn Answer YOUR Recording Requirements

For informative technical data on the basic 1, 2, and 4 channel Sanborn systems, and qualified counsel to help you select the correct Sanborn equipment for your requirements, write to

SANBORN COMPANY

INDUSTRIAL DIVISION
CAMBRIDGE 39, MASSACHUSETTS



IRE People

(Continued from page 50A)

Maxwell K. Goldstein (A'30–SM'46) has been recently appointed President of Balco Research Laboratories, Newark, N. J.

M. K. Goldstein

Dr. Goldstein attended the Johns Hopkins University, from which institution he received his B.S.E.E. degree, and, in 1933, his Doctor of Engineering in Electronics. He was affiliated for many years with the Naval Research Laboratory, and then with the Data Transmission Company as Chief Engineer for Anti-Aircraft Fire Control. He joined Balco Research Laboratories as Vice President.

He has been a member of the technical staff of the Air Navigation Development Board, Department of Defense, and head of Program Research Division, Office of Naval Research. In 1948 he received the Washington Academy of Science Award for distinguished contributions to the Engineering Sciences. He also received the Navy's Distinguished Civilian Service Award for his Direction-Finder contributions to the Battle of the Atlantic.

Dr. Goldstein is a member of Sigma Xi and is listed in American Men of Science.

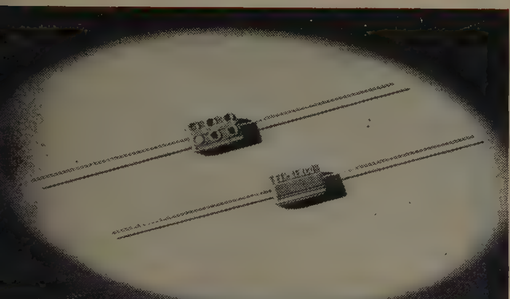
(Continued on page 54A)

trifles make PERFECTION... but PERFECTION is no trifle

#1 IN A SERIES OF TREMENDOUS TRIFLES

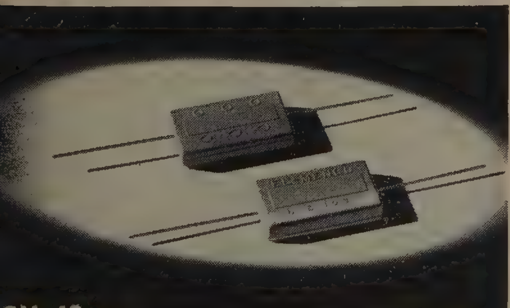
$3\frac{1}{2}$ ounces of perfection

is VITAL to 61 TONS of MAGNIFICENT PERFORMANCE



CM-15

Smallest Molded Mica Capacitors 9/32" x 1/2" x 3/16"



CM-42

Made to Meet All MIL-C-5 Requirements. Largest Molded Mica Capacitors of Wire Terminal Type. 13/16" x 1-1/2" x 5/16"

When the mighty giants of the air lift their massive wings to fly. a thousand and more "tremendous trifles" instantly go to work in harmonious unison to give life and power. It is the perfection of these "trifles" that makes possible the magnificent performance of today's luxurious air liners.

The EL MENCO Capacitor—CM-15—is one of these "tremendous trifles" that plays such a vital part in the efficient operation of aircraft communication.

EL MENCO IS THE ONE OUT OF MANY CHOSEN FIRST

Superiority of manufacture and dependability of performance make EL MENCO first choice on the specification sheet . . . because EL MENCO Capacitors are factory-tested at *double their working voltage* — they are *guaranteed stable* under the most adverse conditions. Whether you use our *high* capacity CM-42 (10-25,000 mmf) or our midget *low* capacity CM-15 (2-525 mmf) you have guaranteed assurance of job-tested, job-rated capacitors — tremendous trifles of perfection so vital to the magnificent performance of YOUR product.

Jobbers and Distributors are requested to write for information to Arco Electronics, Inc., 103 Lafayette St., New York, N. Y. — large stocks on hand — spot shipments for immediate delivery. Sole Agent for Jobbers and Distributors in U. S. and Canada.

ELECTRO MOTIVE is now supplying special silvered mica films for the electronic and communication industries — just send us your specifications.

WRITE FOR FREE SAMPLES AND CATALOG ON YOUR FIRM'S LETTERHEAD

MOLDED MICA **El-Menco** MICA TRIMMER

CAPACITORS

Foreign Electronic Manufacturers Get Information Direct from our Export Dept. at Willimantic, Conn.

THE ELECTRO MOTIVE MFG. CO., INC. WILLIMANTIC, CONNECTICUT

IRE People

(Continued from page 52A)

Isaac L. Auerbach (A'46–M'49–SM'52) has been named Director of the new Special Products Division at the Research Center of the Burroughs Corporation in Philadelphia, Pa.

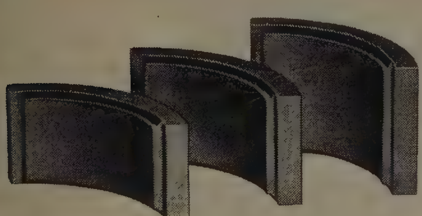
A native of Philadelphia, Mr. Auerbach received his B.S. degree in Electrical Engineering in 1943 from Drexel Institute of Technology. He has done graduate work at both Massachusetts Institute of Technology and Harvard University, from which he received his M.S. degree in Applied Physics in 1947.

From 1943 to 1946 Mr. Auerbach was a Navy radar officer; in 1945 he was assigned to the Naval Research Laboratory for work in IFF systems. Prior to joining Burroughs Research in 1949, he worked on the BINAC and UNIVAC systems at the Eckert-Mauchly Division, Remington-Rand, Inc.

At Burroughs Research Mr. Auerbach has been actively engaged in the design of electronic data-processing systems and has initiated work on magnetic components and circuits; he was responsible for the design of the Static Magnetic Memory System of the ENIAC. He has written numerous technical articles and holds patents in the field of electronic circuits and systems.

Mr. Auerbach is Chairman of the Philadelphia Chapter of the IRE Professional Group on Electronic Computers.

QUALITY FERRITE CORES by ALLEN-BRADLEY



Deflection Yoke Cores in Six Widths

Allen-Bradley yoke cores are all made with an inside radius of 1.027 inches and in six widths as follows: – 0.960 in., 1.125 in., 1.188 in., 1.225 in., 1.250 in., and 1.437 in. Samples for qualification tests can be furnished on request.

The modern Allen-Bradley main plant in Milwaukee, Wisconsin.

UNIFORM IN DIMENSIONS AND PERFORMANCE

The production of Allen-Bradley QUALITY ferrite parts is held to the same close manufacturing standards as all other Allen-Bradley radio and television components . . . such as Bradleyunits, Bradleyometers, and Allen-Bradley ceramic capacitors . . . long recognized as TOP QUALITY. Television equipment manufacturers already consider Allen-Bradley as a desirable source of ferrite cores.

These ferrite cores are now listed in ten part numbers for the U cores and in nine part numbers for the quarter ring cores.

Write for blueprints or samples, today.

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A-B Quality

ALLEN-BRADLEY
RADIO & TELEVISION COMPONENTS
QUALITY

Industrial Engineering Notes

MOBILIZATION*

The Atomic Energy Commission recently released 18 additional patents, including eight in the electronics field. Non-exclusive, royalty-free licenses on the listed patents will be granted by the Commission as part of its program to make nonsecret technological information available for use by industry. Applicants for licenses should apply to the Chief, Patent Branch, Office of the General Counsel, U. S. Atomic Energy Commission, Washington 25, D. C., identifying the subject matter by patent number and title. The following eight patents of interest to the electronics industry were released: Ion Source, 2,668,260; Voltage Regulator, 2,668,272; Relay System, 2,668,934; Electromagnetic Fluid Pump, 2,669,183; Electromagnetic Fluid Pump, 2,669,931; Coupling Stage for Distributed Amplifiers Stages, 2,670,408; Electronic Timing Device, 2,672,556; and Multipactor Tube Oscillator, 2,674,694.

FCC ACTIONS

The Federal Communications Commission recently elected Rosel H. Hyde to continue as Chairman of the FCC until President Eisenhower makes an official appointment. Mr. Hyde's one-year appointment as Chairman expired on April 18. Mr. Hyde was designated acting FCC Chairman by a unanimous vote of the five Commissioners present. FCC rules permit the Commission to choose an Acting Chairman whenever a vacancy exists. . . . **Senator John Bricker, Chairman of the Senate Interstate and Foreign Commerce Committee, recently introduced a bill designed to give the Federal Communications Commission power not only to regulate radio and television stations but networks as well.** Mr. Bricker stated that in view of present network domination in the broadcast field, the ability of an individual station to obtain network programming too often determines whether the station succeeds or fails financially. He also pointed out statistics which show that since the TV freeze was lifted in 1952, 72 TV station grants have been dropped or surrendered by the holders. Of this total, he said, 60 were in the UHF and 12 in the VHF frequency bands. The Senator said he had asked the FCC and other government agencies to submit comments on the bill before the Senators go into the matter further.

TELEVISION

The Federal Communications Commission recently approved a request by the General Teleradio station WOR-TV, New York, to conduct experimental telecasts of the Phonevision system of subscription television for a 90-day period beginning May 15. The Zenith Radio Corp. will install the Phonevision equipment, but there will be no general public demonstrations,

(Continued on page 62A)

* The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of April 26, May 3, 10, and 17, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

Industrial Engineering Notes

(Continued from page 60A)

the original application stated. It was reported in the request that the purpose of the experiments is to finally determine the operating characteristics of Phonevision from a high-powered transmitter in metropolitan New York. . . . **Senator Styles Bridges recently inserted in the appendix of the Congressional Record a copy of a letter he had written to FCC Chairman Rosel H. Hyde, calling his attention to an editorial in the April 30 issue of *TV Guide* which states that only one major set manufacturer so far has specified VHF-UHV tuners for all of its color receivers.** *TV Guide* commented on the difficulties being encountered by UHF stations and stated that, with the one exception noted, other manufacturers contacted "either have made no decision yet or are installing them (VHF-UHF tuners) only as ordered by customers." As a result of this action by Senator Bridges, the Staff Director of the Subcommittee on Communications of the Senate Interstate and Foreign Commerce Committee, which on May 19 will open an investigation of UHF, asked RETMA to poll all set manufacturers and obtain for the hearing information on how many manufacturers plan to install UHF-VHF tuners in their color sets. . . . **In the first case of its kind considered by the Federal Communications Commission—which involves the establishment of a specialized common carrier for the rendition of intercity video relay communication service—the Commission recently granted construction permits to J. E. Belknap & Associates for experimental microwave relay stations using common carrier frequencies to provide an off-the-air pickup and relay of TV broadcast signals for delivery to a community TV system.** The relay stations will be located at Osceola, Kan., and Kennett, Mo., and the signals will be delivered from Memphis, Tenn., to Kennett and Poplar Bluff, Mo. One of the considerations which persuaded the FCC to make a grant of this kind, the Order stated, was that it was "concerned about the expenditures of the individual members of the public who may purchase television sets in reliance upon an expectation of continuity of the contemplated service. We deem it our duty to see that these users are assured of such continuity of service for a reasonable period of time sufficient, at least, to guard against an early obsolescence of their sets. . . . " The Commission also stated that it is not making any express or implied decision as to the existence or extent of any jurisdiction it may have with respect to the installation and operation of any community TV distribution systems. . . . **The nation's fifth educational television station started operating recently when WHA-TV went on the air in Madison, Wis.** The Channel 21 outlet is operated by the Wisconsin State Radio Council in conjunction with the University of Wisconsin. WHA-TV received special temporary authority to start operations from the Federal Communications Commission several weeks ago. It reportedly is operating with power of 1 kw and, according to H. B. McCarty,

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Executive Director of the State Radio Council, will be on the air two hours each evening, and one afternoon a week.

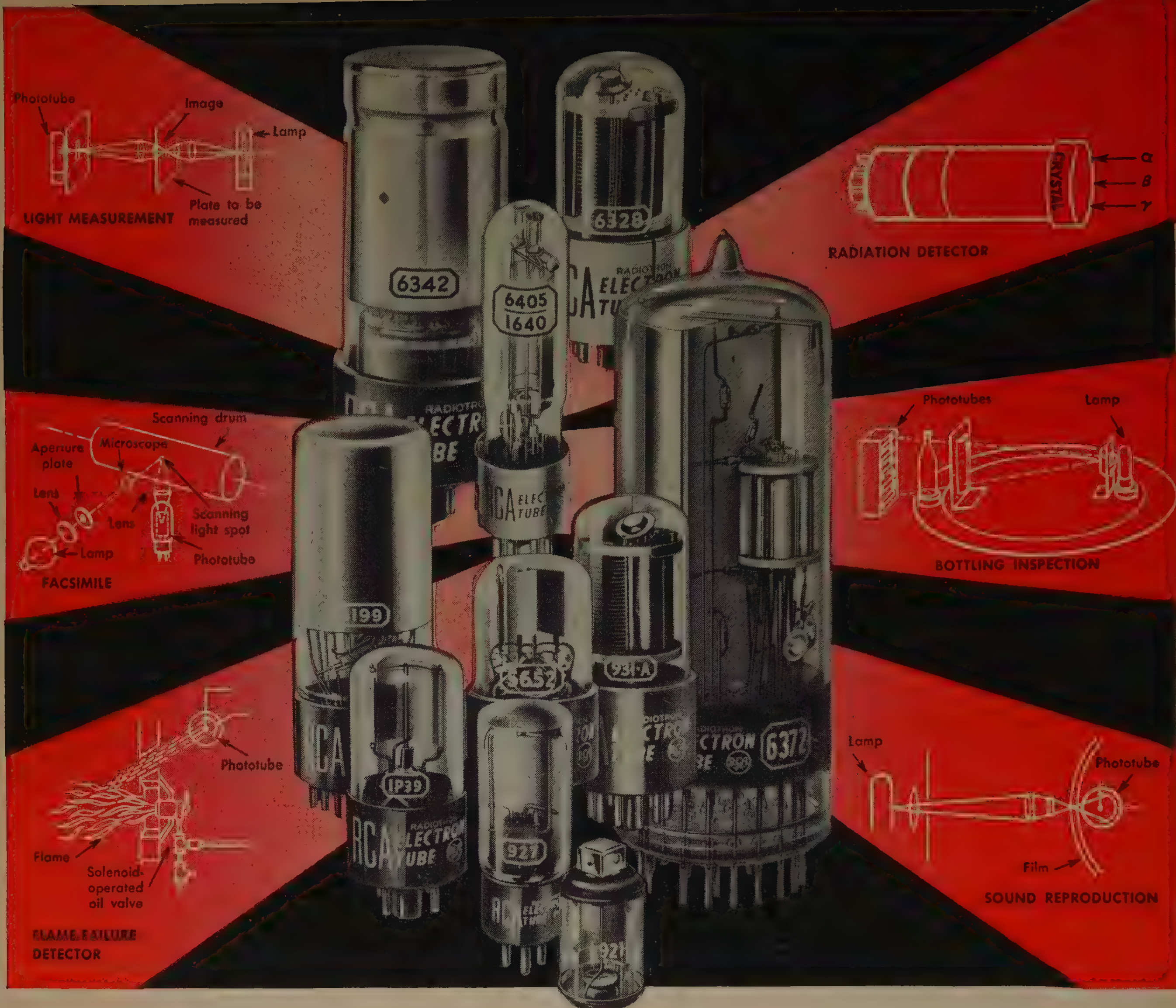
RETMA Activities

The Board of Directors unanimously voted to award the 1954 RETMA Medal of Honor to Chairman Sprague for his contribution to the advancement of the Radio-TV and electronics industry while President and Chairman of the Board during the past four years. The presentation will be made at the industry banquet on June 17 in Chicago at the conclusion of the Association's 30th annual convention. The award to Mr. Sprague was proposed by the Annual Awards Committee headed by Treasurer Leslie F. Muter. . . . After hearing a detailed report of Dr. W. R. G. Baker, Chairman of a Special Committee on Spurious Radiation, on the response of set manufacturers to a RETMA proposal for voluntary submission of TV and FM receivers to an independent testing laboratory for measurement of radiation characteristics, the Board of Directors authorized the Engineering Department to select a testing laboratory and establish operating procedures as promptly as possible. The Board also authorized Dr. Baker to report to the Federal Communications Commission the names of all set manufacturers who agree to adhere to RETMA radiation limitations and the RETMA intermediate frequency engineering standard and who will submit sample receivers to the testing laboratory for certification. . . . **Preliminary plans for the gathering of data on operating experiences with booster and satellite television stations were launched in New York on May 10 at a meeting of the Special Committee of the Broadcast Equipment Section, RETMA Technical Products Division, under Chairman Ben Adler.** With broadcast equipment manufacturers and booster and satellite station representatives present, the group initiated its program which is designed to gather information on these television slave stations for submission to the Federal Communications Commission. Previously, representatives of the FCC had indicated that this data was necessary for the promulgation of rules and regulations for the proposed new service. The committee's work is aimed at extending and improving the broadcast service areas of television stations through the use of boosters and satellites. The group adopted definitions for booster and satellite stations and drew up a long list of problems to be discussed at its next meeting which is tentatively set for Wednesday, June 23. In the meantime, members of the special committee agreed to develop information with the view of solving the problems which will form the basis of RETMAs subsequent recommendations to the FCC.

Technical

The future of automatic production of electronic equipment was discussed at considerable length for some 500 electronic industry representatives and Air Force officials at a "Symposium on Automatic Production of Electronic Equip-

(Continued on page 68A)



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Harry W. Wells

DIRECTOR, 1954–1955

Harry W. Wells was born in Washington, D. C. on January 13, 1907. He received the B.S.E.E. degree in 1928 from the University of Maryland, and the E.E. degree in 1937 from the same institution.

Between 1928 and 1931 he was engaged in radio engineering activities with the Westinghouse Electric and Manufacturing Company, the All-American Expedition to Borneo, and Heintz and Kaufman of San Francisco, Calif. In 1932 he completed pilot training with the Air Force at Kelly Field, Texas, and served on active duty as Communications Officer at Langley Field, Va.

Since 1932, Mr. Wells has been a member of the technical and scientific staff of the Carnegie Institution of Washington, Department of Terrestrial Magnetism. He is Chairman of the Upper Atmospheric Section which has conducted pioneering research in problems of the upper atmosphere, the ionosphere, radio wave propagation, and communications.

More recently Mr. Wells and his associates have pioneered in the study of ionospheric "winds" or traveling disturbances, and have initiated a program of research in radio astronomy for studies of radio frequency radiations from the sun and certain "radio" stars.

Mr. Wells became a member of the IRE in 1936, a Senior Member in 1943, and a Fellow in 1953. He has been active in IRE affairs for many years, including service as officer of the Washington, D. C. Section, the Professional Group on Antennas and Propagation, and several national committees of the IRE. In 1953, he was elected Regional Director Region 3, for the 1954–1955 term.

He has also been affiliated with several committees of the Research and Development Board and served to 1953 as Chairman of Sub-Panel on Propagation, Electronics Committee. In 1947 he received the Award for Scientific Achievement in the Engineering Sciences made by the Washington Academy of Sciences. He is active in the International Scientific Radio Union (URSI) and is Vice-Chairman of its U.S.A. National Committee. His other professional affiliations include the Washington Academy of Sciences, the American Geophysical Union, and the Philosophical Society of Washington.

Education for the Profession

FERDINAND HAMBURGER, JR., FELLOW, IRE

Professor of Electrical Engineering, School of Engineering, Johns Hopkins University

The problems of a young man desiring to enter the engineering profession become more complex as the technological background of the profession continues to expand. What kind of specific preparatory training should he seek? What should be the breadth of his knowledge? How much time should he spend in preparation? What is the importance of it to the man and to the profession?

Either one of two broad educational patterns seems possible—an education based on existing technology and leading to apparently immediate usefulness to an industrial employer; or an education based on broad fundamental knowledge with an apparently longer apprenticeship to achieve equal usefulness in industry. There exists a radical difference in the long-range potentialities of the two forms of education. The first appears short-sighted with long-range potentiality sacrificed for immediacy; the second seems idealistic with immediate usefulness not so evident.

With the pressure to produce, there is a great temptation on the part of industry to expect, and the college graduate to aspire to, immediate productivity. This procedure dictates education at a technological level, and places in jeopardy the future of the profession and its continuing contribution to a fuller life. The true engineering profession lies not only in what we can produce today, but also in what we may produce a decade hence. Such continuing productivity will more certainly come from those trained in fundamental science with an engineering approach than those trained in today's technology.

A secondary aspect of the problem seems equally important. Not only should the young man who trains for the engineering profession be capable of contributing to it, but he should also be educated to appreciate the art of living and be capable of participating fruitfully in the life of the community on a broad scale. This means that his professional educational program should be substantially supported by study in the humanities and social sciences, and he should learn to express himself well by written and spoken word. Therefore, the young man preparing to enter the engineering profession with the desire and ability to make contributions at the professional level can hope, in the usual four years of college, to achieve only a certain facility with the basic sciences of the profession. His continued progress professionally must come from additional study as a graduate student, or continued study on his own initiative.

Except for the social and economic pressures that limit formal college level education to four years in most cases, perhaps the ideal solution would be education over a longer period of time covering broadly both science and technology. Lacking this possibility for the majority of students, both forms of training should continue to be available, but the prospective student should consider his program in the light of both his individual future and the future of his chosen profession.

An Experimental Transistor Personal Broadcast Receiver*

LOY E. BARTON†, SENIOR MEMBER, IRE

Summary—**This paper describes a laboratory-model AM broadcast receiver which used nine alloy-junction transistors and two compensating diodes. Six of the transistors were of the laboratory *p-n-p* type, designed for radio-frequency amplifiers, and three of the transistors were of the conventional *p-n-p* type selected for Class B audio driver and output service. The use of Class B output permitted a total battery drain below 12 milliamperes from six 1.5-volt type C cells in series. The battery life was approximately 500 hours, and the battery cost relatively small. The maximum audio-power output was 150 milliwatts into a four-inch by six-inch oval speaker. The sensitivity and signal-to-noise ratio were comparable to that of conventional battery-operated receivers and conventional ac-dc receivers. The receiver may be used in the place of ac-dc receivers without the necessity of a power outlet or a connecting power cord at a battery cost approximately the same as the cost of power for an ac-dc receiver.**

Introduction

THIS PAPER is primarily concerned with the application of the RCA Laboratories experimental *p-n-p* junction transistors to a broadcast receiver. Early attempts to use the point-contact transistor in a broadcast receiver were discouraging because of low gain, high noise, and comparatively high battery drain. To date these disadvantages were not overcome sufficiently to warrant using the point-contact transistor in a broadcast receiver. The junction transistor, as announced by the Bell Telephone Laboratories near the middle of 1951, overcame the high battery drain objection and to a large extent the noise problem. The frequency response of this new type transistor at that time was limited to a few hundred kilocycles. Early RCA *p-n-p* alloy-type transistors now known as the RCA 2*N*34 junction transistors were also limited in frequency response. However, a small percentage of the RCA transistors had gains up to 20 db at 455 kc. These comparatively few transistors were used to build superheterodyne broadcast receivers. Thus began what was considered a possible practical application of transistors to broadcast receivers.

The alloy-junction transistor was modified to have a lower base resistance and a smaller distance between junctions. These, and other modifications, resulted in much better rf gains for the new experimental transistor. The use of these rf transistors was essential to the performance of the broadcast receiver to be described.

The constructional details of the RCA Laboratories rf transistor have been previously described.[1]

General

The diagram in Fig. 1 shows the complete circuit of the broadcast receiver. The diagram shows a total of nine junction transistors, an oscillator, converter, three IF amplifiers, a second detector, a driver amplifier, and a pair of class B output transistors. The receiver was powered with six type C medium-size flashlight cells. The cells were in series for 9v and a tap at 4.5v was used to permit automatic gain control and temperature compensation, details of which will be given later. The loudspeaker was a 4×6 oval speaker and was selected because its fidelity and sensitivity were considered to be adequate for home use and yet be small enough for normal personal portable-type use.

The improved experimental rf transistors had a low base resistance so that the emitter-input amplifier connection was less subject to regeneration than the base-input connection. A high base resistance would have made the base-input amplifier connection less subject to regeneration. Since neutralization was not planned for the early receivers, the emitter input was used.[2]

The approximate average values for the emitter input characteristics, as listed below, were used in building the receiver IF and rf systems.

	E_c Supply	I_e ma	R_{in}	R_{out}	C_{in} mfd	C_{out} mmfd	DB gain
IF Amp	4.5	0.5	50	50,000	−.006	25	23
Converter	4.5	0.2	100+	100,000	+.006	25	17
Second Detector	9.0	0.2	100		−.006		

A base bias change of about .0025v per degree *C* is needed to maintain constant collector current. To obtain such a bias variation with temperature, a bias from an excessive voltage source was applied to the base and the voltages across resistors R_1, R_2, R_3, R_4, R_5, and R_9 opposed the bias to the associated transistor. This opposing voltage for the individual transistors reduced the net bias for an essentially constant collector current. The value of the collector current depends upon the resistance value of the above individual resistors and upon the voltage of the bias source. This self-biasing system provided temperature compensation and permitted greater interchangeability of transistors. A temperature-compensated bias for the second detector and the class *B* output transistors was obtained from the diode D_1 and will be explained later.

* Decimal classification: R361.116×R282.12. Original manuscript received by the IRE, November 2, 1953; revised manuscript received, February 15, 1954. Also published in the TRANS. I.R.E., P6BTR-5, pp. 6–13; January, 1954.

† RCA Labs., David Sarnoff Research Center, Princeton, N. J.

[1] C. W. Mueller and J. I. Pancove, "A *p-n-p* triode alloy-junction transistor for radio-frequency amplifiers," PROC. I.R.E., vol. 42, pp. 386–391; February, 1954.

[2] A split-input type of neutralization (Patent no. 2,644,859) proved to be quite successful in stabilizing the IF system. When properly used the effective input capacity to the transistor becomes zero at a given base bias.

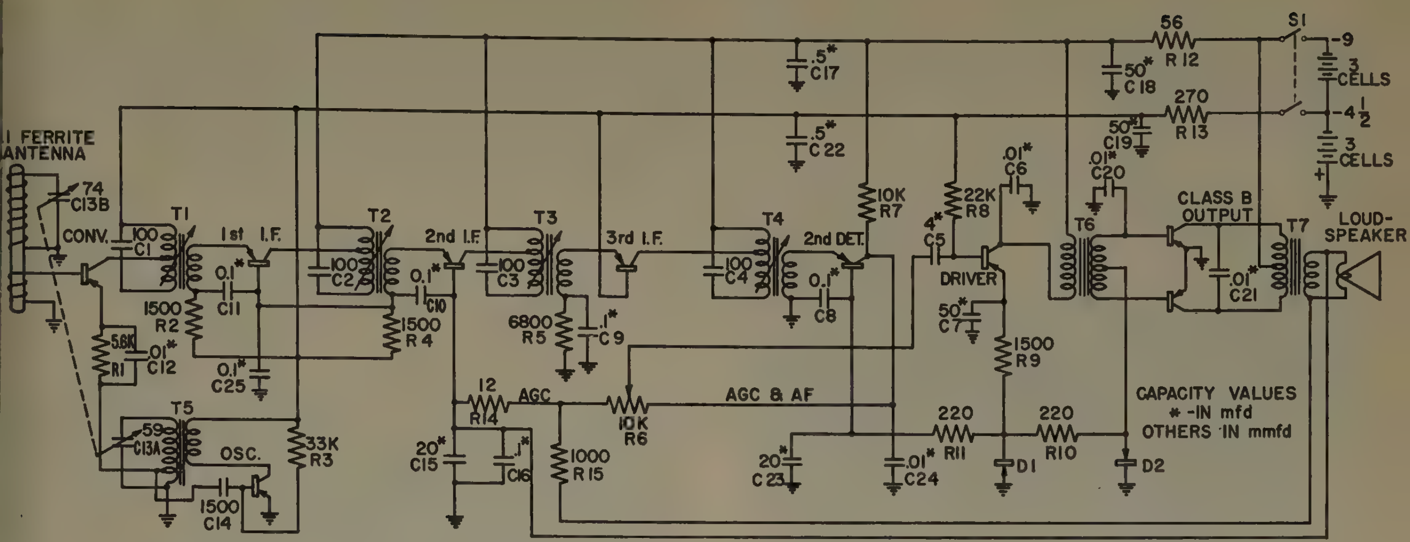

Fig. 1—Circuit diagram for all-transistor broadcast receiver.

RF Circuit Details

The tuned winding of the ferrite-cored antenna of Fig. 1 consisted of about 160 turns of 15/43 Litz wire and had an unloaded Q of about 235. Note that the antenna-tuning capacity is only 74 mmfd, which is less than $\frac{1}{2}$ the value needed for tube receivers. The small capacity was possible because of the low (3 or 4 mmfd) effective shunting capacity of the transistor. The antenna secondary consisted of 4 to 6 turns over the low end of the antenna tuned winding. The output of this winding supplied the signal to the base of the converter transistor. The converter loading of the antenna plus the proximity of the receiver chassis reduced the loop Q to about 100.

The IF transformers were miniaturized for the receiver and will be described in detail later. The oscillator coil was put into one of these transformer assemblies and is shown as T_5. The tuned winding had about 75 turns with a tap at 3 turns above ground for oscillator injection to the emitter of the converter. The tap also served as the feed-back connection for the oscillator. The collector winding consisted of 8 turns tightly coupled to the tuned winding. Here, as was the case for the converter, the effective shunting capacity of the oscillator transistor was about 3 to 4 mmfd which permitted a small tuning capacity. The base bias of the oscillator is arranged for approximately constant current through R_3 from the collector-supply voltage of 4.5v. This constant base current insured that the oscillator start for a wide range of temperatures and compensated for variation in transistors.

Approximately 1v of oscillator voltage was applied to the emitter of the converter through the R_1C_{12} combination across which a dc bias bucking voltage of about 0.7v was generated. This bucking voltage provided temperature compensation for the converter.

The primary of the IF transformers had approximately 150 turns of #38 enameled silk-covered wire. The secondary windings were tightly coupled to the primary and consisted of 3 turns except that the fourth transformer which drives the second detector had a 5-turn secondary. The unloaded Q of the transformers was about 180 and they were loaded to a Q of about 35. A Q of 35 is needed for an IF attenuation of 15 times at 10 kc off resonance for 4-tuned circuits. This attenuation is about normal for tube receivers. The primary tap for the converter was 75 turns above the low rf end of the primary and the corresponding point for the other IF transformers was 50 turns.

The collector voltage for the first and second IF transistors was -9v with the emitters connected through resistor R_2 and R_4 to -4.5 to maintain constant collector current for no signal conditions and resulted in temperature compensation. The resistor R_5 was for a constant collector current and temperature compensation for the third IF transistor.

Automatic Gain Control

The details of the agc are shown in Fig. 1 and may be described as follows. The second IF transistor gets its bias from the second detector collector through the audio volume control R_6 and R_{14}. Since the second detector was biased so that its collector voltage was higher than -4.5v, the base of the second IF transistor was biased negatively and current flowed through its collector and emitter to the -4.5v-battery tap for normal gain. The voltage across R_4 was used for a negative bias for the first IF transistor. The output from the fourth IF transformer was applied to the emitter of the second detector and a signal of a certain magnitude caused the collector current to increase, thus lowering the collector voltage toward the 4.5v point. As this voltage decreased, the bias and thus the gain of the second IF amplifier decreased. The reduction of current in the emitter of the second IF amplifier reduced the bias and therefore reduced the gain of the first IF transistor. This tandem gain-control arrangement placed a minimum dc load on the second de-

tector and also permitted the first IF transistor to decrease its gain more rapidly than the second IF transistor. This sequence of gain reduction is needed to keep distortion to a minimum.

Audio System

A negative audio feedback from the loudspeaker was attenuated by means of the potential divider R_{15} and R_{14}. The voltage across R_{14} was applied to the low end of the volume control R_6 for a degeneration of about 2 or 3 to one at low volume-control settings. This amount of degeneration reduced the distortion from the output of the second detector through the audio amplifier and improved the audio-frequency response of the driver transformer. At maximum volume-control setting for weak signals, the degeneration was essentially zero. However, the volume control was not more than 10 per cent of maximum for most broadcast-station listening.

The audio part of the circuit of Fig. 1 shows a transformer-coupled class B output amplifier which is conventional except for the bias. As is known, the class B amplifier should operate at essentially collector current cut off. To obtain the proper bias for cut off, which was variable with temperature, the current of the driver was passed through its emitter compensating resistor R_9 and through a junction diode D_1 in its low resistance direction to ground. A germanium junction diode having voltage drop of about 0.12v at 2.5 ma at 25 degrees was selected to stabilize the bias against temperature changes. This bias was also the bias needed by the second detector and, therefore, was applied to the base of the second detector through R_{11}. The resistance R_{11} must be kept low to prevent dc degeneration in the detector which would interfere with the agc action. Bias for the class B output transistors was taken from the diode D_1 through an isolation resistor R_{10}. A diode D_2 was also needed to provide a path for the peaks of audio current to the bases of the class B transistors. Without D_2 the peak current to the bases of the output transistors may exceed the current through D_1 and the resistance of the base circuit to ground would become very high, resulting in clipping of the audio output. These peak currents would also affect the bias of the second detector and the agc action if D_2 were not used.

As a matter of interest the ratio of transformer T_6 was 5 to 1 to each side of the secondary. The available peak signal-input current to the base of the output transistor would then be 5 to 10 ma depending upon the excitation current required by the driver transformer. The diode D_2 was required to take the peak currents. The output transformer T_7 loaded each output transistor during the time it functioned to about 180 ohms. Under these conditions the power output was about 150 mw and the peak collector currents were about 45 ma.

Dynamic transfer curves for two selected transistors are shown in Fig. 2 which indicate the bias required at room temperature. At about −25 degrees C. the curves shift to the broken line curves and indicate the necessity for a device for controlling the bias as the temperature changes. The diode D_1 controls this bias to the proper value because the germanium-junction diode had essentially the same temperature characteristics as the germanium-junction transistor.

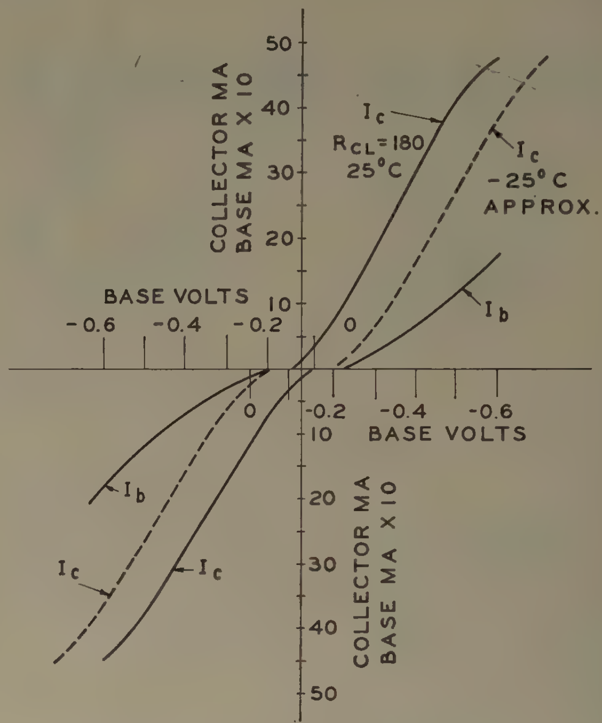

Fig. 2—Class B dynamic transfer curves.

IF Transformers

The only major item requiring miniaturization was the IF transformer. These transformers were tuned impedance-matching devices and the coils were totally enclosed in ferrite.

The first design consisted of a spool of ferrite material on which the coil was wound. This spool was inserted into a ferrite cylinder and a ferrite disc was moved toward or away from one end of the spool for tuning. The leads were taken through notches at the other end of the spool. The ferrite pieces were appropriately mounted in metal sleeves for chassis mounting and tuning adjustments. This transformer measured $\frac{1}{2}$ inch high and $\frac{1}{2}$ inch in diameter.

A second IF transformer was designed which had two ferrite cups with cores extending from the bottom to the open end of the cups. A coil was placed over the core of one cup and extended over the core of the second cup when placed face to face with the first cup. The ends of the cores were slightly tapered so that tuning was effected by turning one cup with respect to the other. The leads were brought out through a slot in the bottom cup. The two cups were held together by pressure when assembled in an appropriate chassis mounting can. The second transformer was about $\frac{7}{16}$ inch high by about $\frac{3}{8}$ inch in diameter. The unloaded Q of either type was approximately 180.

Other components for the receiver were commercially available except the tuning condenser which was made especially for the receiver. The frame of the tuning condenser, because of the low capacity values needed, was only $\frac{7}{8}$ inch deep and the oscillator plates were cut for 455 kc. The tuning range was 535 kc to 1650 kc.

Complete Receiver

The complete receiver chassis is shown in Fig. 3 and shows the general arrangement of components and the relative size of the speaker. Another view as mounted in a transparent case is shown in Fig. 4.

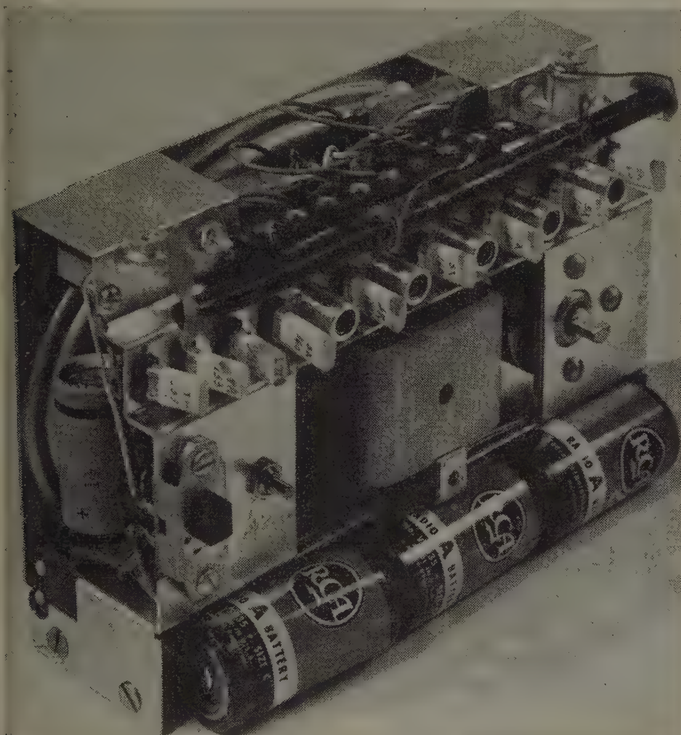

Fig. 3—Transistor receiver, less case.

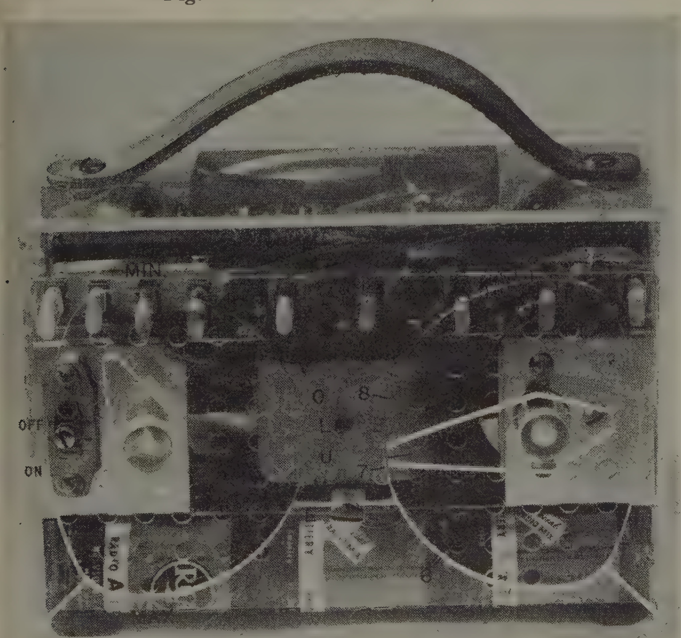

Fig. 4—Front view of experimental transistor receiver.

A pocket version of the receiver is shown in Fig. 5 and is identical to the receiver described except for battery and speaker size. The speaker is a 3-inch-round type and the battery consists of 6 penlite cells. The chassis shown in Fig. 3 was mounted in an opaque case and is shown as the lower receiver in Fig. 5. The cabinet or case for the respective receivers was functional with no particular emphasis on styling.

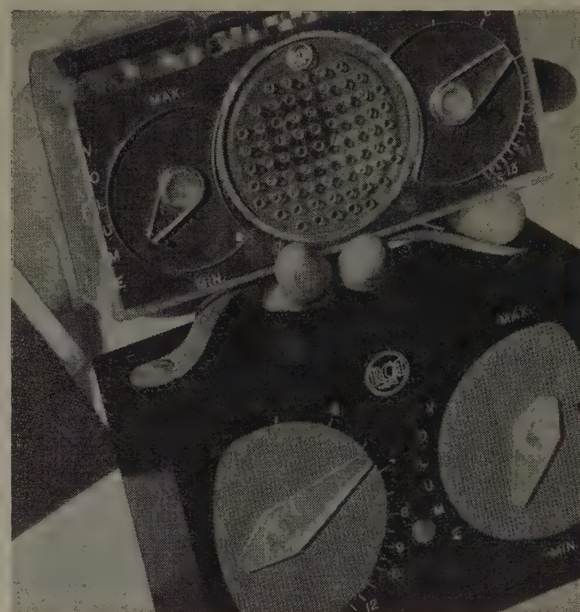

Fig. 5—Pocket and personal transistor receiver.

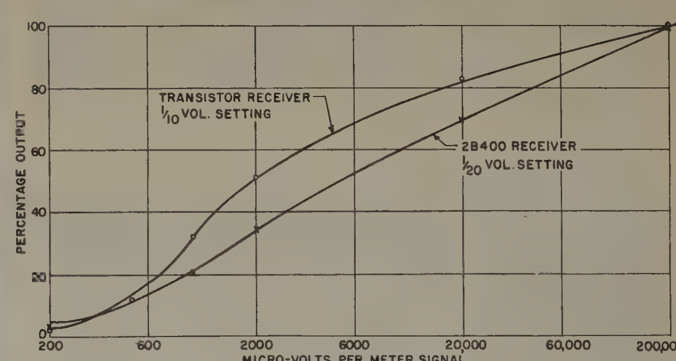

Fig. 6—Agc characteristics.

Receiver Performance and Data

The performance of the agc is shown in Fig. 6. A corresponding curve for a battery-operated tube receiver is shown for comparison. Note that the transistor receiver has a flatter agc.

The over-all audio response is shown in Fig. 7 for a resistance load. The volume control was at less than 10 per cent maximum in this case so that near maximum audio feedback was used. A similar curve is shown for a battery-operated tube receiver.

A signal-to-noise curve vs rf input at 1,000 kc is shown in Fig. 8. This is close to the best performance achieved to date. Measurements of the same type made on a personal battery receiver gave a 5-mw sensitivity

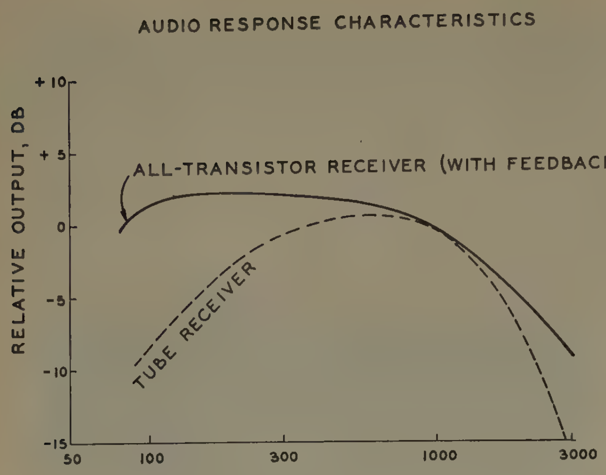

Fig. 7—Audio frequency, cps.

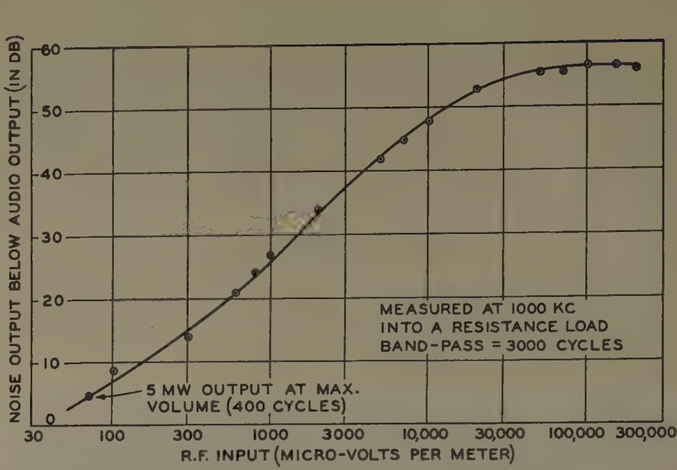

Fig. 8—Noise characteristic of a transistor receiver.

at 85 microvolts per meter and about 2-db less signal-to-noise ratio to about 1,500-microvolts input. At higher inputs the signal-to-noise ratio was 3 to 4 db higher for the transistor receiver.

The performance and characteristics of the two transistor receivers as compared to a representative battery personal tube receiver are tabulated in Table I. The pocket receiver could be reduced in size by using a shorter ferrite-cored loop antenna, a smaller speaker, and special batteries, but this sacrifice in performance seemed unjustified at this time because the receiver easily goes into a trouser or coat pocket.

TABLE I

Comparison of Transistor Receiver with Battery-Operated Tube Receiver

	Tube (Personal)	Transistor (Pocket)	Transistor (Personal)
Sensitivity (5mw out.) μv/m	80	40	40
Noise Ensi (μv/m)	14	14–17	14–17
Maximum Audio Output mw	75	150	150
Loudspeaker Size inches	2×3	3 dia.	4×6
Weight with Batteries	3.9 lbs.	1.4 lbs.	2.75 lbs.
Over-all size inches	9×6×2¾	6½×3¾×1½	6½×4½×2½
Over-all Volume cubic inch	140	33	73
Total Battery Power mw	920	100	100
Batteries Used	Special	6 penlite	6 med. flashlight
Battery Life hours	80–100	50	500
Battery Cost per Hour cents	5.0	1.2	0.15

The audio volume, fidelity, and economy of operation of the larger transistor receiver was comparable to the usual ac-dc table-model receiver. Therefore, it could be used in the place of the power-line receiver without the disadvantage of an attached power cord. The cost of batteries to operate the receiver was about the same as the cost of power for the average ac-dc receiver.

The receiver also operated quite well in a car when it was coupled to the car antenna by means of an rf lead connected to a ferrite-cored inductance which in turn was placed in the vicinity of the receiver loop. This car arrangement eliminates the shielding effects of the car body and also eliminates the directivity of the receiver. The receiver may then be mounted at any desired place in the car within the limitation of car noise and removed for portable and home use.

Acknowledgment

The writer wishes to acknowledge the co-operation of C. W. Mueller of the RCA Laboratories, and his associates, for providing experimental transistors having the necessary performance to build the above receiver. The writer also wishes to thank Chandler Wentworth of the RCA Laboratories for his part in the design and construction of the IF transformers.

CBS Television City Technical Facilities*

H. A. CHINN†, FELLOW, IRE, R. S. O'BRIEN†, SENIOR MEMBER, IRE, R. B. MONROE†, SENIOR MEMBER, IRE AND P. E. FISH†, SENIOR MEMBER, IRE

Summary—**The Columbia Broadcasting System recently inaugurated television program service from its new Hollywood Television City headquarters. The engineering and architectural design of this modern television broadcasting plant is based upon a flexible and expansible master plan that provides for the ultimate construction of 24 studio units.**

The initial construction phase now completed provides two audience and two nonaudience studios, each exceeding one-quarter acre in area, together with the necessary technical, production, scenery construction, and office facilities to completely support CBS-Hollywood television operations.

This paper describes the philosophy underlying the design of the video, audio, and communication facilities for this project. Emphasis is placed on description of methods and features that are new or novel.

Introduction

CBS TELEVISION CITY, in Hollywood, California, is a fully self-contained and expansible television production center designed expressly for the efficient production of television programs. The initial construction, shown in Fig. 1, provides a total floor area of more than eight and one-half acres and consists of two adjoining structures, viz., a studio building and a service building. The studio building houses four one-quarter acre studios, three rehearsal halls, technical facilities areas, reverberation chambers, dressing rooms, and a half-acre area for scenery and property storage. The service building houses scenery construction shops, maintenance shops, production offices, and public reception areas.

The interior arrangement of the studio level of the initial buildings is shown in Fig. 2, page 1068. The four studios, two of which accommodate studio audiences, project outward from the service building with a 40-foot wide central set-trucking corridor providing direct, level traffic flow between receiving platform, scenery and craft shops, and studios. A factory-like flow of materials through the various shops to the different studios and thence, by way of freight elevators, to storage in the basement has been achieved.

In addition to scenery and property storage, the basement level (Fig. 3, page 1068) contains dressing rooms, primary power installations, mechanical building facilities, maintenance shops, and the heart of the audio and video system—the central technical area.

The initial construction of Television City is but a small fraction of the ultimate plant planned for the 25-acre site. Additional studios may be provided by extending the two initial studio wings or by adding additional ones on both sides of the central service building. A total of as many as 24 one-quarter acre studios may be provided in this manner. The service building may be expanded proportionately in height, as well as in length, to keep pace with studio expansion, and office buildings may be constructed along the perimeter of the property as required.

The functional architectural design has resulted in a television production plant that is complete and well integrated in its initial phase, but one specifically planned for ease and flexibility in expansion. The basic concept of providing a complete but readily expansible installation has also been followed in the design of audio and video facilities for the plant. Together with the large physical extent of Television City, this philosophy has led to the introduction of many innovations in system planning and component design. This paper describes the engineering aspects of these innovations and the audio and video facilities in which they are incorporated.

Fig. 1—CBS Television City's initial unit: two adjoining buildings providing $8\frac{1}{2}$ acres of floor space. The studio building extending to the right houses four one-quarter acre studios; the three-story glass-walled service building at left is devoted to scenery shops, maintenance shops, and offices for creative and production personnel.

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† CBS-Television, Engineering Dept., New York, N. Y.

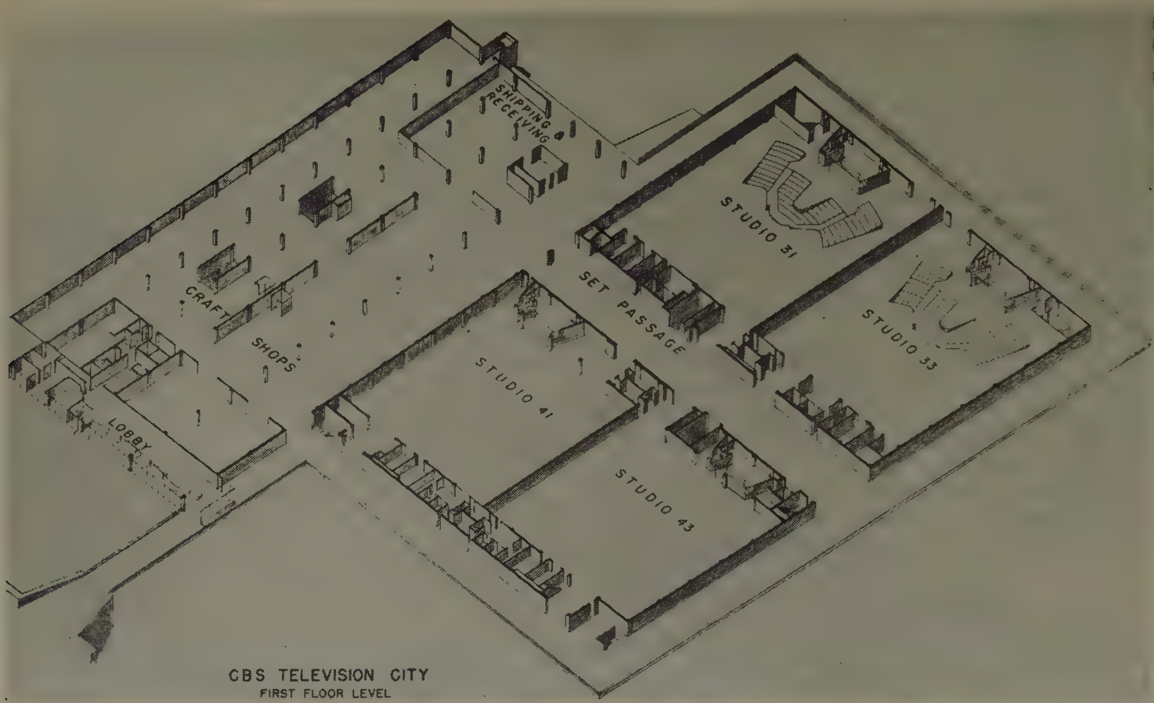

Fig. 2—This sketch of the studio floor level shows the easy accessibility from receiving platform and craft shops to studios by means of 40-foot wide set passage. Vehicles can enter building by ramp at receiving platform and drive into any studio. Studios 31 and 33 have an audience seating section.

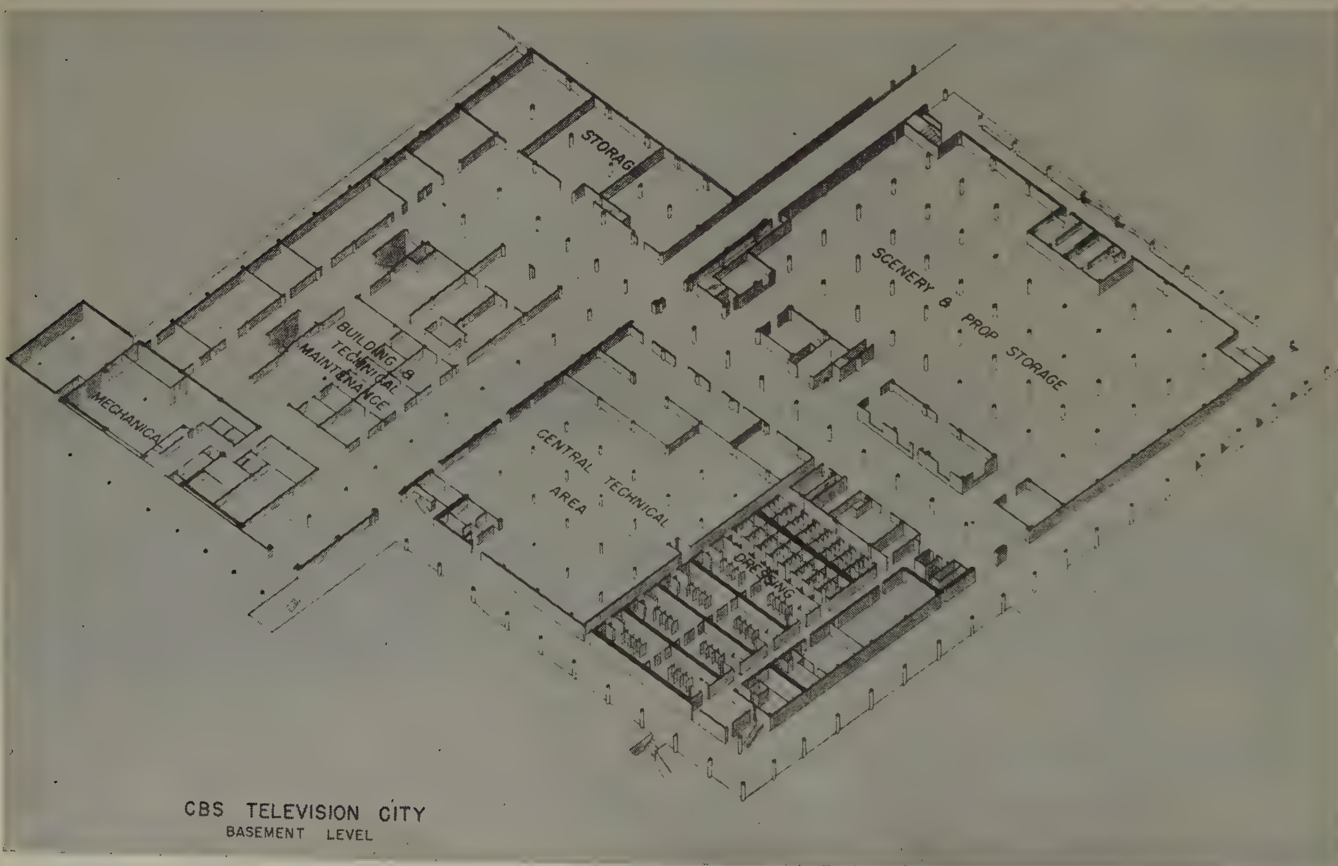

Fig. 3—The basement level is utilized for the central technical area, dressing rooms, maintenance shops, and storage. Over one-half acre of space is provided for the storage of scenery and properties.

General System Considerations

In a large television studio plant the audio and video facilities are, perforce, separated into functional groups located at various points throughout the plant. In Television City the facilities grouped in these technical areas are interrelated as shown in Fig. 4.

The facilities that serve all studios are located in a central technical area on the ground level under the

studio floor near the geographical center of the ultimate plant. The central technical area is subdivided into master control, telecine (a contraction of television-cinema), program control, and television recording areas.

Master control is the location of both the first and the last elements of the video system, viz., the centralized pulse generation and distribution facilities essential for synchronous scanning in all cameras, and the program trunk switching and distribution facilities. Terminal facilities for program transmission to, and reception from, off-premise locations are included. In the parallel audio system, master control provides similar facilities except that the first element does not exist.

Telecine is the central source of audio and video signals obtained from recorded program material such as motion picture film, still picture slides, or other video and audio recording media. Such recorded material is employed in television broadcasting to provide complete television programs, portions of programs, or short program sequences. This material is usually transmitted from the telecine room to a studio control room where the recorded material is integrated into live portions of the program.

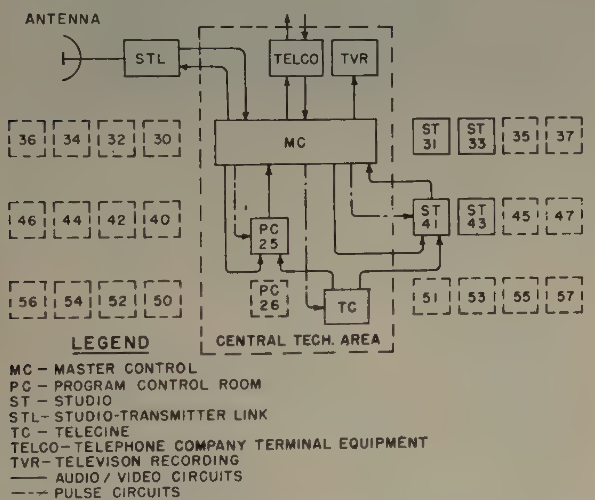

Fig. 4—This simplified block diagram shows the general functional relationship between a typical studio and the technical facilities in other areas of Television City. Facilities serving all studios have been located in the central technical area; facilities associated with a single studio have been decentralized and located in the control room of the studio. Note that numbers of future studios (dotted boxes) have been pre-assigned to insure an orderly numbering system regardless of the sequence of expansion moves.

Program control rooms are equipped with switching and mixing facilities identical to those in a studio control room, but no cameras or studio area are associated with them. They serve the same function as a studio control room when telecine program material or program material from a remote point is being handled. Program control rooms are also used when it is desired to route a studio program through a secondary control point wherein alternate regional announcements may be inserted or other program editing accomplished.

Television recording, which is also provided for in the central technical area, contains the equipment necessary for making 16 mm and 35 mm picture and sound recordings of television programs.

Contrasted with the centralized equipment, those audio and video facilities which serve only a particular studio are decentralized and located in the control room of that studio. It has long been considered good practice in aural broadcasting to make each studio audio installation an essentially self-contained unit. However, in video systems for television studio plants, a large portion of the video switching equipment for each studio has often been located in a master control or central equipment room because of video-signal timing considerations.

In Television City, a timing compensation system has been evolved whereby the complete studio video switching system can be located in the studio control room. This important innovation makes each studio an essentially self-contained production unit from the viewpoint of both audio and video facilities; it also provides complete flexibility in the choice of monochrome or color television equipment in present or future studios. Decentralization of studio facilities makes the technical space required for each studio an integral part of the studio, greatly reduces potential space requirements in a central area, and alleviates the problem of providing such space far in advance of the ultimate plant expansion.

The remaining technical areas, shown in Fig. 4, are concerned with transmission of signals to and from various off-premise locations. Such transmission is handled for the most part by the telephone company whose cable terminal facilities are in the master control room. For certain remote program applications, or under emergency conditions, the cable facilities may be supplemented by microwave radio links. Microwave equipment is housed in a transmitter-receiver room in a fourth-floor penthouse.

Supplementing the technical areas are a large maintenance shop with an adjoining garage for two mobile equipment trucks, film viewing and cutting rooms, and a number of monitor-equipped conference rooms and offices.

Studios

The audio and video installation in all existing Television City studios is essentially identical. However, since some additional facilities are required to serve a studio audience, the most comprehensive account of studio facilities can be made by describing one of the audience-type studios (see Fig. 5 on page 1070).

Studio audio and video facilities are in the studio control room which is located to the rear of the audience, against the rear wall of the studio. In the control-room plan view (Fig. 6, page 1070) it can be seen that the control room personnel are grouped around the program director, who sits at the left end of the director's console. Program assistants are at his right and key technical per-

Fig. 5—Interior view of an audience studio shows the wide expanse of useful stage area and the close relationship of studio audience to the stage. The entrance to the control room is at the left. The control console in the foreground controls over 300 kilowatts of incandescent studio lighting through a bank of thyraton tubes. A memory circuit permits immediate return to any one of five predetermined settings of the 60 dimmer circuits by operation of a simple pushbutton selector switch.

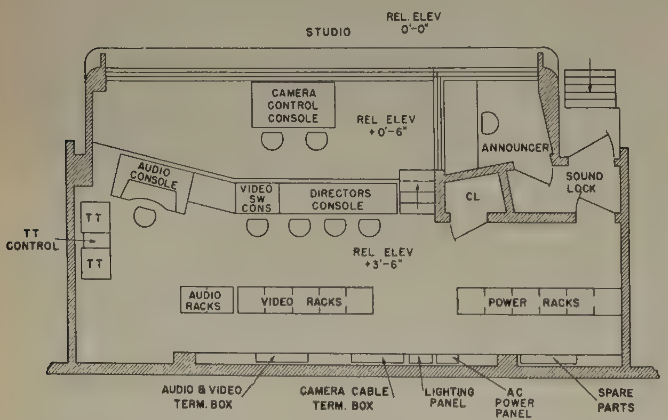

Fig. 6—Plan view of typical studio control room. Program director is seated at left end of directors console; technical director at the video switching console.

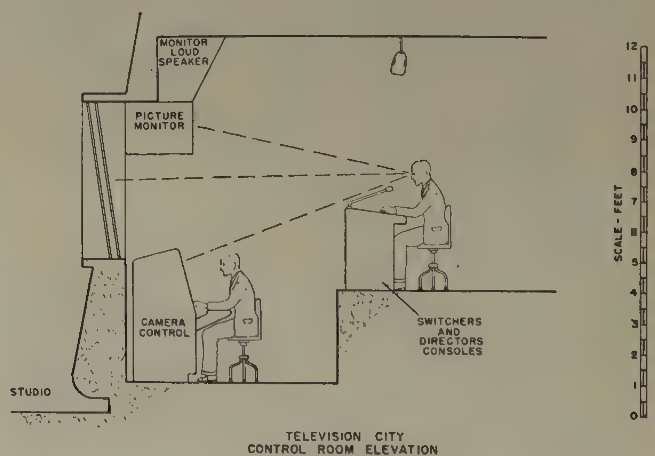

Fig. 7—This control room section shows the excellent visibility of both the control room monitors and the studio areas from the upper level. This arrangement permits all concerned with picture quality to make observations on the same high-quality picture monitors.

sonnel at his left. The technical director, at his immediate left, operates the video switching console and is responsible for technical phases of program production. The audio operating position is far enough to the left that a glass partition may be erected around this section to give a measure of acoustical isolation should this be desired. The audio console has been set at a slight angle to enhance the line of sight between audio operator, program director, and announcer.

The control room utilizes a two-level arrangement in which the camera-control console is on the lower level, a three-foot difference in height insuring that the technical director, program director, and others on the upper level have an unobstructed view of the camera-control picture monitors over the heads of the camera-control operators. This arrangement, detailed in Fig. 7, is very desirable as it enables all the persons who are concerned

Fig. 8.—Upper level of a studio control room. At extreme lower left is the turntable control unit with its two associated transcription turntables. Right of the turntables is the audio control console, a utility desk, video switching console, and the directors console. The picture monitor to the left is for the audio operator; the grouping of picture monitors to the right includes two preview monitors, a cue monitor, and an illuminated clock. Above the monitors is a loudspeaker baffle. A second loudspeaker baffle, upon which are mounted the two monitoring loudspeakers for the audio-control operator, can be seen directly above the audio console. The lower level of the control room is not visible.

with picture quality to make their observations on the identical high-quality camera-control picture monitors.

The announcer, in a glass-walled announce booth, has a clear and unobstructed view into both the control room and studio. He is seated in a position from which he can easily receive visual cue signals, if necessary, from either the audio operator or the director.

Auxiliary control-room equipment is mounted in air-conditioned equipment racks which form the rear boundary of the control-room operating area. Audio and video terminal blocks for cable runs to both the studio and the central technical area are mounted in a wall cabinet on the rear wall of the control room.

A. Studio Audio Facilities

Audio facilities in the Television City studios center around a CBS 3E audio control console (Fig. 8). Basic considerations involved in planning and designing similar audio consoles have already been described[1,2] and will not be repeated here. Instead, this description will be concerned mainly with the over-all audio and communication system and particularly with features and engineering developments that are new and different.

1A. *Physical Design Considerations.* In the CBS Television City studios the audio console has been supplemented with additional equipment units. These include: (*a*) a turntable control unit, (*b*) an equipment rack containing facilities required for studio sound reinforcement and for the studio audience reaction microphones used only in audience studios, and (*c*) a second rack containing complete studio communication facilities.

This division of equipment results in an extremely flexible audio system as the various units can be used alone, or in combination, to equip both aural and television broadcasting studios of various types. For example, the audio console alone provides all program audio facilities normally required in an aural or television broadcasting studio originating large-scale productions. If, in addition to the extensive facilities provided by the console, transcription reproducing facilities are also needed, the turntable control unit with its two associated transcription turntables may be added. Further-

[1] H. A. Chinn, "CBS studio control-console and control room design," Proc. I.R.E., vol. 34, p. 287; May, 1946.

[2] R. B. Monroe and C. A. Palmquist, "Modern design features of CBS studio audio facilities," Proc. I.R.E., vol. 36, p. 778; June, 1948.

more, if the studio accommodates an audience, the rack containing audience equipment may be added. Finally, if the studio involved is for the production of television programs, the rack containing television studio communication equipment may be added. Thus, it can be seen that the division of equipment that has been employed permits either an aural or television broadcasting studio to be equipped in the most economical manner as only those facilities actually required for the intended operation need be installed.

Physical considerations influencing the design of the CBS 3E audio control console include: (*a*) functional location of all controls and all visual monitoring facilities, (*b*) excellent visibility from operating location to studio and to other parts of the control room, (*c*) ready access to all parts for maintenance, (*d*) adequate ventilation, and (*e*) an attractive appearance.

Audio design features that have proven very successful in earlier consoles and which have been included throughout the entire Television City plant include: (*a*) single plugs and jacks, because of their smaller size and superior characteristics,[3] (*b*) use of only two audio amplifier types, both plug-in, and two audio tube types, to simplify maintenance, (*c*) similar standardization of other components to reduce to a minimum the number of types to be stocked for plant maintenance, and (*d*) standardization on RETMA standard circuit impedance of 150 ohms to assist in achieving uniform response-frequency characteristics at the higher audio frequencies.

Other features of earlier audio consoles that have also been retained include: (*a*) a built-in lectern for the program script, (*b*) illuminated scale markings on the mixer control knobs that also serve as channel-in-use indicators, (*c*) multi-colored single-window pilot lights to achieve a clean-cut uncluttered front panel appearance, and (*d*) "magic-slate" mixer channel designation-pad and control knob setting indicator associated with each mixer control to eliminate pencil marking of panels.

The turntable control unit is self-contained, including within its small dimensions all associated amplifiers, power supply, cueing loudspeaker, and jackfield. It is mounted to the left of the audio operator between the two transcription turntables and, in this position, the controls are within his reach for normal operation. However, on unusually complex programs, the physical separation of console and turntable control unit makes it possible to employ a second man to operate the turntables without interfering with the normal operation of the console.

2A. *Mixer Circuits.* A simplified block diagram, Fig. 9, indicates the scope of the audio facilities provided in the Television City studios. These facilities are capable of simultaneously mixing 21 of the following 28 program sources which are normally connected to the console, viz., (*a*) eleven studio microphones, (*b*) six audience reaction microphones, (*c*) one announcer's microphone, (*d*) two transcription turntables, (*e*) one sound effects console, (*f*) four film projectors, (*g*) two remote lines, and (*h*) one reverberation line.

Each microphone channel includes a plug-in type preliminary amplifier to amplify the extremely low microphone output voltage prior to mixing. On the other hand, signals from other sources, including the turntable control unit, reach the console at a sufficiently high level that preamplification is unnecessary. After passing through a bridged-T mixer control, the signal on each channel is divided by means of a multi-output resistance isolation network which features a high degree of isolation between each of its outputs. Microphone channels one through six are divided three ways, one of the outputs feeding the regular program channel, the second output feeding the studio sound reinforcement channel, and the third feeding the reverberation channel. The remaining channels are not normally used for reverberation effects and the networks on these channels, therefore, provide only two outputs.

A four-position microphone sub-mixer has been included in the console. The output of this sub-mixer may be routed through any one of the regular microphone mixing channels by means of a single patchcord connection. A portion of a program, e.g., a multi-microphone orchestra pickup, may be pre-balanced by means of this sub-mixer and this balanced portion of the program then gained by means of a single mixer control which would become, in effect, a sub-master volume control. Sound reinforcement outputs are not provided from the individual sub-mixer channels as the feed associated with the channel through which the sub-mixer is connected is available if sound reinforcement is required.

Facilities are included in audience studios for mixing the outputs of the six audience reaction microphones. As in the case of the sub-mixer, the output of this audience-reaction mixing circuit is patched into one of the regular microphone mixing channels. Volume controls are not required in the individual audience microphone channels but it will be noted that a key switch has been included in each channel. These switches permit silencing of any audience microphone should this prove necessary.

The "*A*" and "*B*" mixer channels handle audio from film projectors and from remote lines. Program material to these two channels may be selected from any one of six sources by means of electrically and mechanically interlocked pushbutton selectors. Four of the pushbutton positions have been assigned to audio feeds from the telecine patchcross unit (see below) and the two remaining positions to remote lines which are fed from master control. Other sources of program may readily be connected to any of these six positions by means of a patchcord connection at the console.

Individual mixer-channel key switches have not been employed as their omission promotes more pleasing audio control since each microphone must of necessity

[3] H. A. Chinn and R. B. Monroe, "Single jacks for broadcast application," *Audio Eng.*, vol. 31, p. 12; July, 1947.

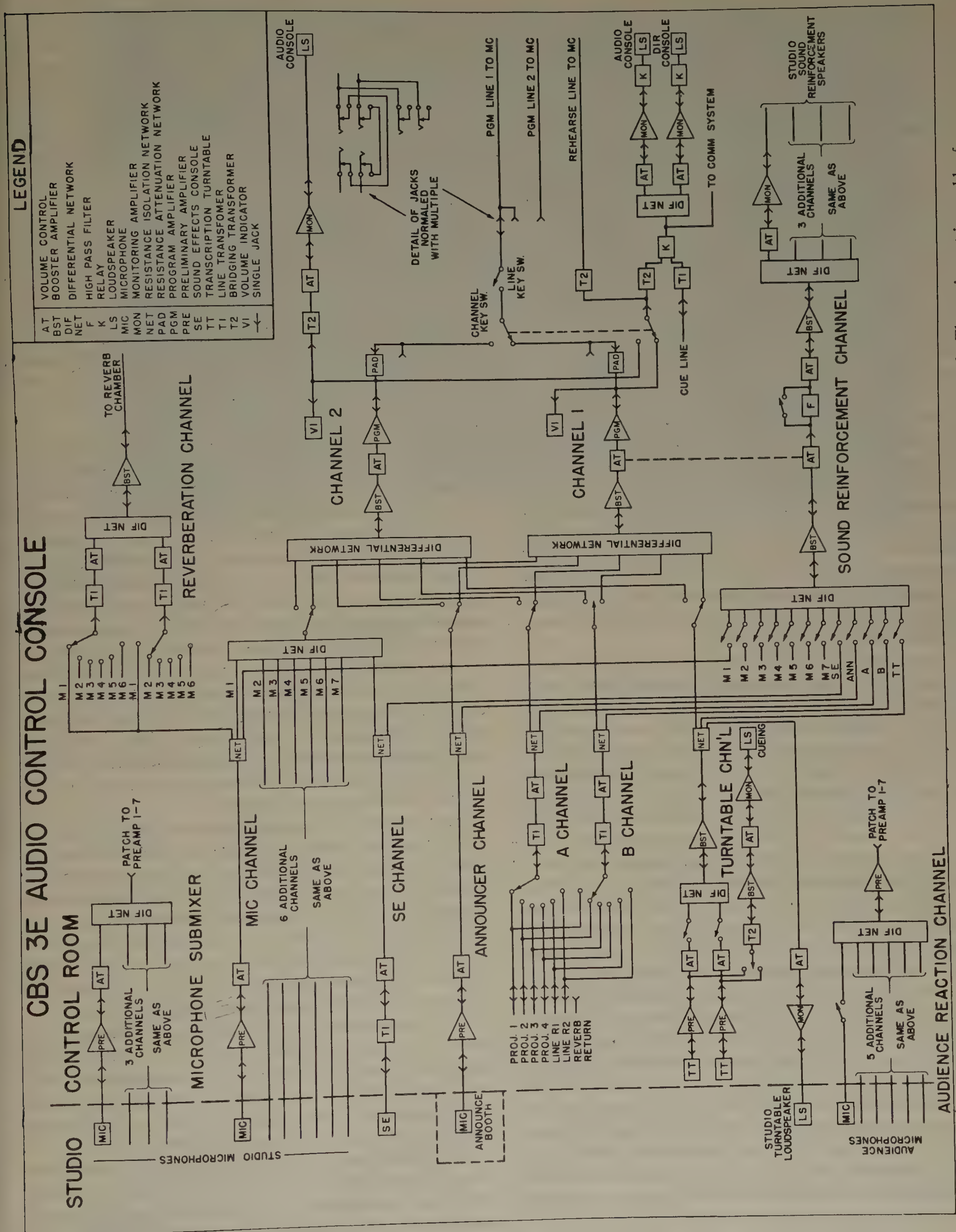

Fig. 9—This simplified block diagram indicates the scope of facilities provided by the studio audio console. The equipment is capable of simultaneously mixing 21 of 28 normally connected program sources.

be faded in and out rather than abruptly turned on and off.[2] All regular studio microphone channels, together with the sound effects channel, are controlled by a single key switch. Other key switches control the announcers microphone channel, the turntable channel, the "*A*" channel, and the "*B*" channel.

3A. *Program Channels.* Two program channels have been provided in the CBS 3E console. Program material from the mixer key switches is connected to channel 1 by throwing the desired switch down, and to channel 2 by throwing the switch up. In the center position, program material and the input to both channels are terminated. This switching arrangement requires the use of a separate five-position differential network at the input of each of the two channels as well as an eight-position differential network for combining the seven studio microphone and sound-effect channels.

Channel 1, which is the regular and normally used channel, contains a booster amplifier, master gain control, program amplifier, volume indicator, and line pad. Channel 2 is identical to channel 1, including not only a separate volume indicator but a complete and separate monitoring channel. Either of the two program channels may be connected to the outgoing program line by means of the channel key switch. The regular aural monitoring facilities, which feed monitoring loudspeakers both at the audio console and at the directors console, are bridged across the channel that has been selected by the channel key switch.

The second program channel serves several purposes. Primarily, it is an emergency channel that can instantly replace the regular channel in the event of a failure in the regular channel. When not so used, it serves as a test channel for checking quality and transmission level of any portion of a program prior to switching it to the regular channel and, in this use, is similar to the preview monitoring channels provided in the video switching system. It is also available as a utility microphone-to-line or microphone-to-loudspeaker channel for use in obtaining special or unusual program effects.

4A. *Turntable Control Unit.* The self-contained turntable control unit provides facilities for mixing program material from two low-level transcription reproducing turntables. In addition to the audio output to the audio console, the turntable unit also provides one high-level audio output to a self-contained cueing loudspeaker (used for setting the stylus to the desired starting point on a disc recording) and another high-level output to a studio loudspeaker. The latter loudspeaker is used when it is necessary for performers in the studio to hear transcribed program material as it is being broadcast.

5A. *Sound Reinforcement.* Television or aural broadcasting studio sound reinforcement requires a different mixing technique than that employed in preparing program material for the usual broadcast purpose.[4] Effective sound reinforcement requires that only those portions of the program that cannot otherwise be easily heard by the studio audience be reproduced on the sound reinforcing loudspeakers. For example, an orchestra can easily be heard throughout the entire studio and reinforcement of the microphones making this pick-up would only tend to jumble the performance reaching the studio audience. However, a soloist performing with the orchestra may require reinforcement to be easily heard in all parts of the audience area.

In the past, a second mixing console and operator, often located in the audience section of the studio, has sometimes been employed for the specific purpose of preparing the program material for sound reinforcement and properly controlling the level of reinforcement during the performance. A different method of sound reinforcement has been employed in the Television City studios which does not require either this second mixing console or the services of the second operator.

As shown in Fig. 9, a program sample for sound-reinforcement purposes is obtained from the resistance isolation network following the mixer volume control in each mixer channel. The sound-reinforcement signals are brought to key switches which determine whether or not the particular channel will be reinforced. The output of all selected channels is combined in a differential network. The output of this network is amplified, then passes through a volume control which is ganged with the program-channel master volume control, through a filter that rolls off the low frequencies, a sound-reinforcement master volume control, and on to amplifiers that feed the studio loudspeakers. With this arrangement, all adjustments made in gaining the regular program material, whether on the mixer or master volume controls, are also equally effective in controlling the gain of the sound-reinforcing channel.

Operation of this system is simple. Microphones, or other program sources that are to be reinforced, are selected as required by means of the sound-reinforcing selector key switches. The level of reinforcement is established by means of the sound-reinforcement master volume control. The low-frequency roll-off filter, which permits a higher level of reinforcement before the point of singing is reached, can be switched in or out of the circuit as desired. No other operations are necessary.

6A. *Reverberation Facilities.* In the production of television programs, especially those of a dramatic nature, program sequences are often encountered where it is necessary to create the illusion of greater-than-normal reverberation. These greater-than-normal reverberation effects are achieved in the following manner.

A program sample for reverberation purposes is obtained from the same network in each mixer channel that supplies the sound-reinforcement signals. Program samples from microphone channels one through six are brought to two pushbutton selector switches which permit either one or two microphones to be selected for reverberation effect at any one time. If reverberation is

[4] H. A. Chinn and R. B. Monroe, "Broadcasting studio sound reinforcement," *Audio Eng.*, vol. 31, p. 5; December, 1947.

Fig. 10—The sound-effects station in each studio is located on a balcony overlooking the studio staging area. Right of the sound-effects console is a studio floor monitor, a flexible assembly of equipment that combines in a single unit, a picture monitor, loudspeaker, clock, and interphone station.

being applied to two channels, volume controls permit different degrees of reverberation to be applied to each of the signals.

The microphone channel or channels selected for reverberation are amplified and transmitted to a reverberation chamber, or to a synthetic reverberation device, via master control. The output of the reverberation device returns to the studio on the reverberation return line where it is introduced into the program channel by means of the *A* or *B* mixer channels.

The desired reverberation effect is obtained by adjustment of the relative proportions of the original and the reverberated signal which, in turn, is controlled by the relative settings of the microphone channel mixer control and the reverberation-return mixer control.

7A. *Utility Facilities.* Although not shown in the block diagram of Fig. 9, the CBS 3E audio system includes a full complement of utility facilities which may be connected into the circuit by means of patchcords whenever some special operating requirement arises. These utility facilities include a sound-effects filter which is usually connected into a microphone channel when it is desired to produce an effect requiring a restricted bandwidth, as well as a dialogue equalizer which is connected into an audio channel when it is desired to adjust the audio perspective to conform to the visual program material.

Other utility facilities include a utility volume control (which consists of a bridged-T attenuator equipped with isolation transformers on both input and output in order that it may be used with either balanced-to-ground or one-side-grounded circuits), two utility double-pole triple-throw key switches, a utility bridging transformer, and seven sets of four parallel-connected jacks.

8A. *Studio Sound-Effects Facilities.* Sound effects play an important role in the production of television programs. The sound-effects station in the Television City studios (Fig. 10) is located on a balcony overlooking the studio and has been equipped with a CBS 4A sound-effects console[5] which provides in a single, compact, mobile unit all audio facilities normally required to handle the sound-effects requirements of the most complex television programs.

One of the most valuable sources of sound-effects material is the large library of sound-effects recordings that is available. The sound-effects operator must be equipped with facilities to reproduce these recordings and, in addition, he must also be equipped to handle sounds originating from other sources where the output may be either a sound wave or an audio signal.

The CBS 4A sound-effects console includes three variable-speed turntables and four record-reproducing arms. The mounting arrangement that has been employed enables two reproducing arms to be used with each of the three turntables, a feature that permits continuous sound to be obtained from a record and also makes possible echo effects by proper placement of two styli in the record groove. A low- and high-frequency equalizer is provided in each of the four record-reproducing channels.

[5] R. B. Monroe and P. E. Fish, "CBS-TV sound effects console," *Audio Eng.*, vol. 34, p. V12; March, 1950 and vol. 34, p. V12; May, 1950.

The variable speed turntables and equalizers increase the scope and expand the usefulness of sound-effects recordings by permitting the sounds to be altered and modified to suit the particular requirement at hand.

In addition to the mixing positions for the four turntable channels the sound-effects console also provides two utility mixing positions which accommodate sound-effects microphones, additional disc or magnetic tape reproducers, audio from special sound-effect devices, or audio from remote lines.

Two outputs are provided from the console, one at line level, the other at high level for loudspeaker operation. A standard volume indicator permits the operator to maintain a uniform output line level.

Physical design features of the sound-effects console include (*a*) aluminum housing to reduce weight, (*b*) aluminum roll-top cover to protect equipment when not in use, (*c*) transparent plastic script and record holder which does not obstruct forward vision of operator, (*d*) grain-of-wheat lamps in each pickup arm to aid in record cueing, and (*e*) color coding to associate pickup arms with their corresponding mixer knobs.

In operation, the complete sound-effects portion of the program is transmitted to the control-room audio console at standard line level. At the same time, the high-level output of the sound-effects console feeds a studio loudspeaker whereby the performers hear the sound effects. It is the usual practice to locate a sound-effects loudspeaker at each set in the studio. A push-button selector on the control panel of the console permits the operator to switch the high-level audio to the loudspeaker at the set in use.

B. Studio Communication System

Most visitors to properly designed television studios are impressed by the excellent co-ordination existing between the large group of people involved in the production of a program. This co-ordination results from the use of extensive systems of studio communications whereby each person involved is kept informed, and is cued and directed in his activities.

The service that a studio communication system must render is the distribution of intelligence to all engaged in the production of the program in a manner such that the activities of the entire group of technical, production, and performing personnel are properly co-ordinated.[6] The intelligence necessary for program co-ordination takes the following forms; (*a*) *spoken words*, such as direct instructions from the program or technical director, (*b*) *audio program material* from which many cues are obtained, e.g., the orchestra leader usually receives cues for his music from the dialogue of the program, (*c*) *picture program material* which provides information to almost all concerned with the program, e.g., the sound-effects operator for synchronizing his sounds with studio action, the audio operator to avoid boom-suspended microphones from appearing in the picture, and the lighting staff to check the studio lighting.

[6] R. B. Monroe, "CBS television studio intercommunication facilities," *Audio Eng.*, vol. 34, p. V3; December, 1950.

Aural communication is accomplished by means of five basic systems: (*a*) an interphone system for telephonic communication, (*b*) a headphone cueing system, (*c*) an induction-field transmission channel, (*d*) a loudspeaker talkback and call system, and (*e*) a two-way intercom with telecine. Picture monitors are built into most operating locations. In addition, portable studio floor monitoring units (Fig. 10), which combine in a single unit a picture monitor, loudspeaker, interphone station, and a clock are used wherever needed, e.g., at locations on the studio floor where the performers and other studio personnel can see the production exactly as it is leaving the studio.

1B. *Interphone System.* The lower portion of Fig. 11 shows the studio interphone system employed in the CBS Television City studios. As can be seen, a private telephone circuit is provided between each studio camera and its associated control. However, when a switch at the camera-control position is closed, this interphone circuit is connected to a conference bus and the cameraman and camera-control operator may then communicate with the program director, technical director, or any other camera or station also connected to the bus.

In normal operation, the camera interphone circuits are connected to the conference bus. However, the private camera to camera-control circuit is essential. For example, in the alignment of a camera chain it is possible for cameraman and camera-control operator to communicate without interfering with, or being distracted by, other conversations on the system.

Television cameras are usually equipped by the manufacturer with two built-in interphone stations. In CBS studios, one of these stations is used by the camerman and the second by the cameraman's assistant, who is employed when a dolly type of camera mounting is used. Each of the camera's two interphone stations are equipped with a twin jack arrangement whereby one jack is associated with the interphone circuit and the other is associated with a separate cueing circuit intended for the transmission of a second source of information (e.g., audio program material) to a second headphone unit. Most CBS cameramen prefer to use headsets with a single receiver unit (connected to the interphone) as this leaves one ear free to hear directly the sounds in the studio. On a few types of programs, particularly those with a large orchestra, the intensity of the direct studio sounds make it difficult to hear the interphone. Under these circumstances, a headset with two receiver units is used. It will be noted that the arrangement employed makes it possible to connect either audio program material or the voices of the program and technical director to the second receiver unit by means of a single patchcord connection in the control room.

An amplifier-reinforced feed from the technical and program director's desk microphone circuit is introduced into the interphone conference bus. This permits all interphone stations connected across this bus to hear the director when he speaks even though he may not be using the microphone portion of an interphone headset. The reinforced output of the desk microphone is ad-

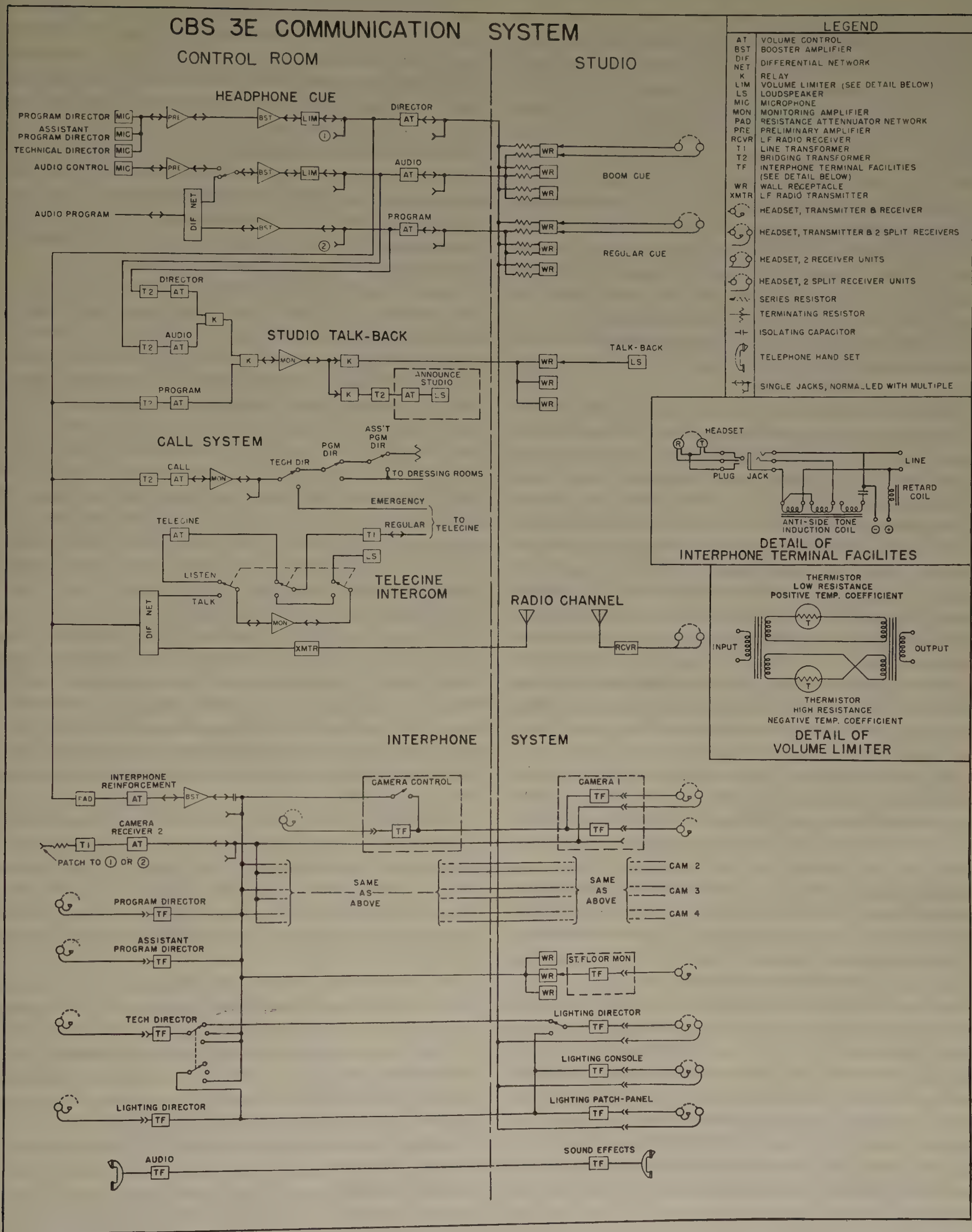

Fig. 11—This block diagram shows the extent of the communication system provided in each studio. Simple volume limiters in each microphone channel maintain the audio levels substantially constant.

justed slightly higher in level than other interphone conversations, thus giving the directors an appropriate degree of priority over other conversations on the interphone system.

A second interphone system serves the studio lighting personnel. There are stations in the studio for the lighting director, lighting-control console operator, and lighting patch panel operator. In addition, a second station is

provided for the lighting director in the control room should program conditions require his presence there. The technical director has been provided with a switch that permits the following circuit alignments: (a) private conversation between technical director and lighting director, (b) interconnection of lighting and main interphone conference buses, and (c) isolation of the lighting interphone from the main interphone system.

Cases frequently occur where an interphone station is required in the staging area of the studio, e.g., an orchestra leader may require two-way communication with the control room. Inasmuch as studio monitors, shown in Fig. 10, are usually available on the studio floor, an interphone station has been included in each of these units. This interphone station can be connected to the interphone conference bus by means of a patchcord connection in the control room.

The audio operator in the control room frequently must work out special problems with the sound-effects operator in the studio. For this reason, a private interphone circuit has been provided between these two.

The interphone system constitutes only one portion of the studio communication facilities and is augmented by the additional circuits also shown in Fig. 11 to provide the communication means necessary to serve all personnel concerned with the studio production. As this block diagram shows, three different channels of information, the output of the director's microphones (which includes the technical director, program director, and the assistant program director), the output of the audio operator's microphone, and audio program material are amplified and routed to various headphone, loudspeaker, and transmitter circuits.

2B. *Volume Limiter.* In the past, one of the major shortcomings of studio communication systems has been the wide difference in volume level between different users of the system. If the gain of the system was adjusted for the average person, soft-spoken ones had difficulty being heard, while others with louder voices, or those having a tendency to speak close to the microphone, would overload the system. The obvious solution to this problem is a volume limiter and this has been included in the Television City facilities.

As already mentioned, in the interest of simplifying maintenance of the plant, only two audio amplifier types have been employed. Volume limiting was accomplished, therefore, not by a volume-limiting amplifier which would have introduced another amplifier type, but by a device[7] employing thermistors which maintains a uniform volume level throughout the system the output of all microphones in the communication system (except the carbon-button transmitters in the interphone system) pass through one of these limiting devices.

3B. *Headphone Cueing Circuits.* It is necessary to provide communication circuits to persons in the studio not served by the interphone system, such as microphone boom operators, announcers, and sound-effects operators. This is accomplished by means of conventional two-receiver headphones which have been wired to permit the reproduction of different information in each ear. These split-headphone cueing facilities are known as ***headphone cue.***

Headphone cue circuits are one-way, that is, a microphone is not provided for talking back to the control room as is the case with the interphone system. In practice, however, two-way communication with the control room is possible during studio rehearsals by means of the regular studio microphones.

Two different types of headphone cue are used in CBS studios. The first type is for the general use of technical and production personnel and is known as ***regular*** headphone cue. Regular cue reproduces the voices of the program and technical director in one of the two earphones and audio program material in the other earphone. The second type of headphone cue is specifically for the use of the microphone boom operators and, for that reason, is known as ***boom*** headphone cue. Like regular cue, boom cue reproduces the voices of the program and technical director in one earphone and audio program material in the other; however, the control-room audio operator can break into the audio program side of the circuit and talk directly to the microphone boom operators at any time.

4B. *Induction-Field Communication Channel.* Some persons involved in the production of television programs, notably the studio floor manager, must be free to move to any part of the studio without the encumbrance of a cable, yet must remain at all times within the range of the voice of the director in the control room. An induction-field communication channel is provided for these persons.

The system employed in the Television City studios is an amplitude-modulated induction-field system which operates with a power of a few watts in the low-frequency range between 100 and 175 kilocycles. Different frequencies are assigned to adjacent studios. The receiver is an ultra-compact unit which is carried over the shoulder.

5B. *Studio Loudspeaker Communication.* Communication to performers and others in the studio is accomplished during studio rehearsals by means of talkback loudspeakers in a manner similar to that used in aural broadcasting studios. Control-room talkback facilities are provided for the program director, assistant program director, technical director, and audio operator.

Inasmuch as television studios are very large compared to aural broadcasting studios, it has proven convenient to employ a small, mobile talkback loudspeaker unit at each of the sets in use. The loudspeaker in the studio floor monitor is usually employed for this purpose.

6B. *Telecine Intercom.* Because of the physical separation of the telecine room and studio, a two-way loudspeaker intercom has been provided between the two points. Via the telecine patchcross unit (see below), this intercom circuit is extended directly to intercom sta-

[7] J. A. Weller, "A volume limiter for leased-line service," *Bell Lab. Rec.*, vol. XXIII, p. 72; March, 1945.

tions at each of the projectors assigned to the studio, as well as to the stations at the associated camera control units. An emergency intercom circuit has also been provided between control room and telecine for use in the event of a failure of the regular circuit, or in the event that communication is desired with telecine before the necessary patchcross connections have been made.

7B. *Call System.* A loudspeaker call system has been provided from each studio control room to the dressing rooms. This call system permits the program director, or his assistant, to call each performer to the studio well in advance of his scheduled appearance. A loudspeaker has been installed in each of 24 dressing rooms as well as in the associated corridors and in other places where performers may congregate. It will be noted that the same amplifier is utilized for this service as is employed for the telecine emergency intercom circuit. The key-switch wiring gives the telecine emergency circuit priority over the dressing-room call circuit.

C. *Studio Video Facilities*

Features of the studio video installation in the Television City studios include: a flexible dual-fader video switching system, full remote control of telecine projectors, an extensive studio monitor-bus system, uniquely-supported audience picture monitors, separately-powered emergency program signal channels, full use of matched-impedance video circuits, and generous allocation of expansion space to accommodate future monochrome or color television requirements.

1C. *Video Switching System.* The video switching system in Fig. 12, shown on page 1080, is divided into two sections, the first section handling noncomposite (i.e., picture plus blanking) video signals, the second, handling composite (i.e., picture plus blanking and synchronizing pulses) signals.

The noncomposite section provides input positions for picture signals from four telecine cameras, from four studio cameras (currently image-orthicon type), and from three spare input positions for future cameras. A twelfth input position is fed from the output of the "effects" channel mixer amplifier, thus making it possible to either switch or lap-dissolve into this channel by means of the faders incorporated in the "line" channel mixer amplifier. This makes possible the smooth introduction of pre-set superimpositions or other effects. The video signal on any one of the noncomposite input positions, including the effects position, can be previewed at any time on either of the two preview channels. The video signal from the "line" mixer amplifier is applied to an isolation amplifier group wherein synchronizing pulses are added, the resulting composite signal being fed into the second, or composite, section of the video switching system.

There are four inputs to the composite signal switching section. One is the video signal from the noncomposite switching section. The other three are input positions for remote or other composite signals which are routed to the studio from master control. Like the noncomposite signals, any of the video inputs to the composite switching section can also be previewed on either of the preview channels. Provisions have been made to remove local synchronizing pulses automatically from the monitoring circuits when a composite signal is selected for previewing.

Although the over-all video system is planned to permit direct switching of video in each studio control room, video switching relays and bias-controlled fader (or mixer) amplifiers were used in the initial installation because of their more reliable and more easily maintained performance as compared to direct switching devices available at the time the installation was designed. The relays and amplifiers are located in the equipment racks immediately behind the technical director and are controlled from the video switching console that he operates.

The similarity between the functional block diagram and the video switching panel layout can be noted by comparing Figs. 12 and 13, pages 1080 and 1081. On the panel, the lower two rows of internally illuminated, self-indicating[8] momentary contact pushbuttons and the levers, to their right, control the most frequently used "line" channel; the middle two rows and associated levers control the "effects" channel, and the upper two rows control the two "preview" channels. Beyond the fader levers, to the right, are the composite section pushbuttons for both the program and preview channels.

The pushbuttons that select the four input channels normally associated with telecine signals are separated slightly from the other noncomposite inputs and are located at the left where they are conveniently adjacent to the corresponding projector-control pushbuttons. Projector remote control circuits are arranged to start and stop any telecine projector assigned to the studio or to turn the projector lamp on without film motion when it is desired to observe the continuous projection of a single frame of film.

Just above the grouping of projector control buttons are the studio monitor-bus selector buttons which control video relays in one of the studio equipment racks. As shown in Fig. 12, three monitor buses are provided, each of which can be independently switched to the studio output signal, a program cue signal from master control, a patchable utility input, or to either of the two preview channels. Picture monitors at various points throughout the studio may be patched to any of these three busses as dictated by a particular operating setup. A typical assignment is illustrated in Fig. 12.

2C. *Audience Picture Monitors.* Several of the audience picture monitors can be seen in Fig. 14. They are enclosed in special hooded housings suspended from adjustable spring-loaded pantographs. They can easily be positioned in elevation and angle to give the studio audience an optimum view of the program as it is being picked up by the studio cameras. Equally important,

[8] Self-indicating video switcher push buttons, originally introduced in the CBS-Television Grand Central studios in 1948, have become commonplace in custom installations.

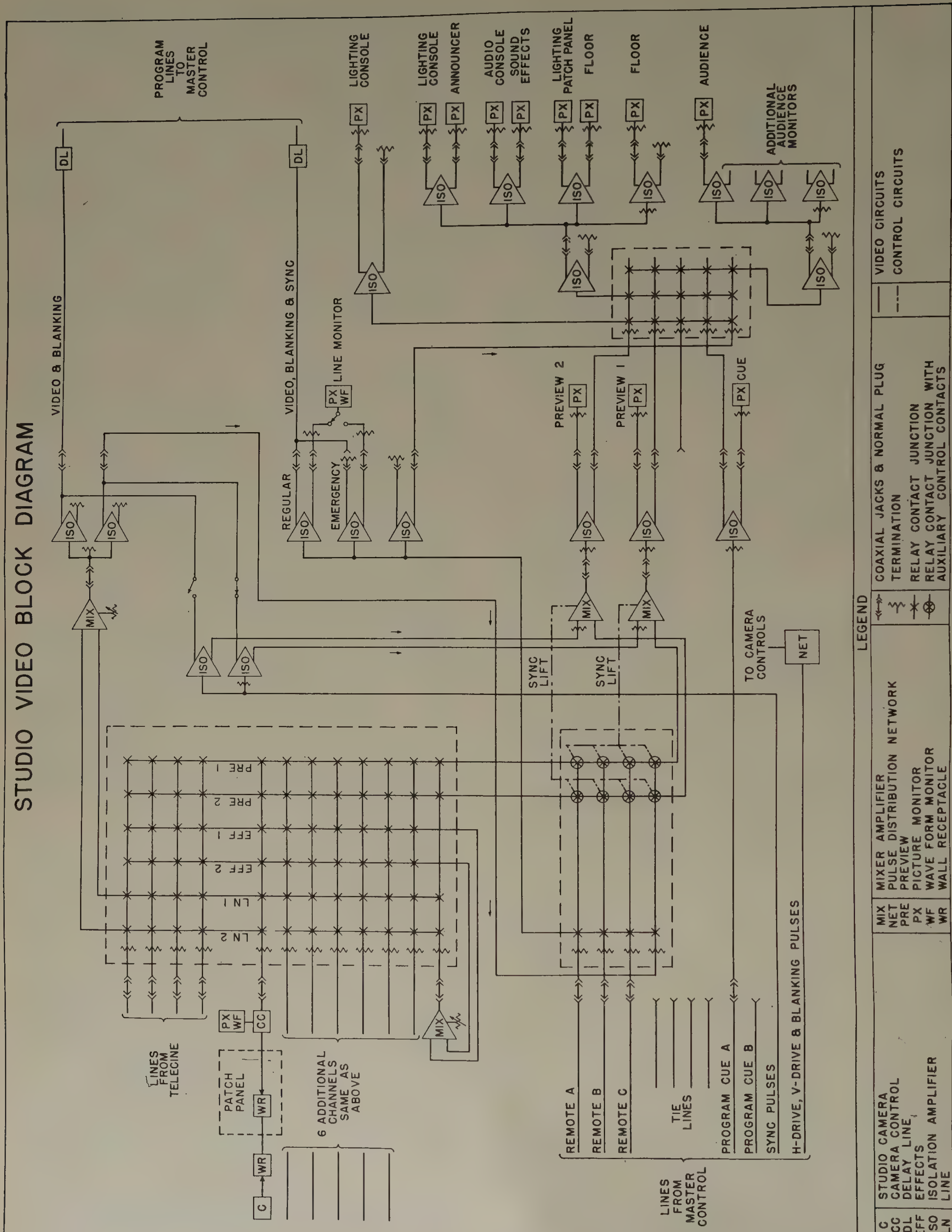

Fig. 12—This simplified block diagram shows the flexibility of the studio video facilities.

these monitors permit the audience to see portions of the program that do not originate in the studio, such as film or remote sequences.

3C. *Emergency Facilities.* One incidental advantage accruing from the decentralization of studio video facilities is the reduction in time required to localize and remedy trouble in the video system. To further expedite the process of keeping a program on the air, extensive emergency provisions have been included. As can be seen by referring to Fig. 12, should trouble develop in the "line" channel, the "effects" output may be patched directly to the sync-mixing isolation amplifier group. The power supplied to the "effects" channel is completely separate, even to the extent of being fed from a different ac phase from that powering the "line" channel. Similarly, the two parallel amplifiers where synchronizing and video signals are mixed, are fed from separate power sources.

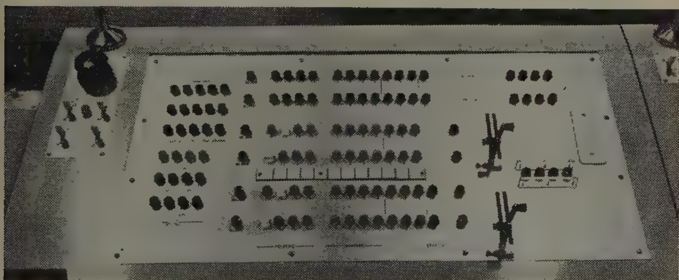

Fig. 13—A close-up of the studio video switching console. Illuminated pushbuttons indicate activation of a particular circuit. The small panel to the left contains intercom controls and a monitoring loudspeaker volume control.

The regular and emergency output amplifiers associated with the composite switching section are also operated from separate power sources. Monitor feeds are supplied from each of these amplifiers to the control-room line monitor in the camera-control console. The emergency monitor feed can be checked at any time, but the nonlocking type switch employed on this circuit keeps the monitor normally connected to the regular monitor feed.

The studio output channel and line that normally feeds noncomposite studio signals to master control (and thence to a program control room when inter-studio co-ordinated operation is required) also serves as an emergency studio output channel. If necessary, synchronizing pulses can be added to this channel by the operation of a switch.

The video jackfields, where emergency and operational patches are made, are located, together with all amplifier units and video switching system components, on the equipment racks which establish the rear boundary of the control-room operating area.

4C. *General.* In the design of the studio video system, emphasis has been placed upon flexibility, reliability, and the avoidance of complex circuits and components. Thus, synchronizing signals are added in an ordinary additive mixer using standard commercial amplifiers. Clamping is applied in the simplest possible way at only

Fig. 14—Picture monitors serving the studio audience are suspended from spring-loaded pantographs which support the monitor at any desired elevation from three to 15 feet above the stage floor level.

two points. Almost all of the video circuits have been arranged to operate on a matched impedance basis and cable equalization has been used, as described below, where needed to insure high quality performance. A pulse cable termination circuit, in which the camera-control input capacitances are lumped into a shunt capacitive element of a bridged-T network, is used to obtain reflection-free timing pulses.

Telecine

Program material from motion picture film constitutes a large portion of a typical broadcasting schedule and is used, at least in short sequences, to augment the live portions of almost every television studio program. However, in the course of the rehearsal and broadcast of a program, the film facilities are operated for much shorter periods of time than the studio facilities. This fact, coupled with the complexity and high initial cost of the projection equipment, dictates a central location from which audio and video signals derived from film or other recording media may be distributed to all studios.

The operations performed in the telecine projection and control rooms are in the nature of preparation and incidental adjustment of signals. The actual operations required to make a film insertion in a studio program are performed, properly, in the studio control room where all other aspects of the production are controlled and directed.

A. Design Considerations

The telecine area in Television City (Fig. 15) is divided into a projection room and a control room. In the initial installation three 35 mm projectors, three 16 mm projectors, and two opaque still picture projectors are provided together with their associated cameras. Expansion space is available to double this complement.

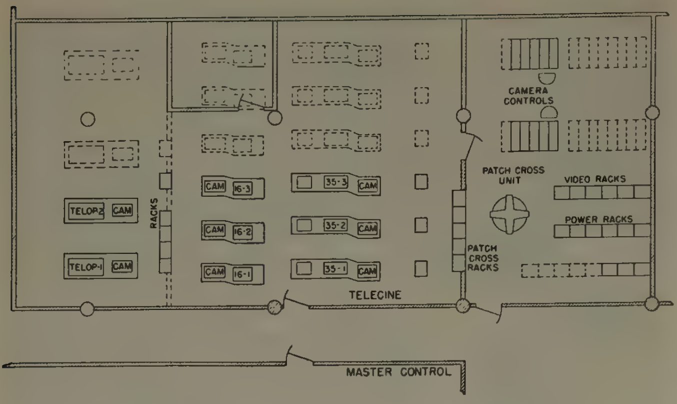

Fig. 15—Plan view of the telecine area which has been divided into two sections: projection room on the left, control room on the right. Future expansion is indicated by dashed lines.

The adjacent telecine control room houses the control and monitoring units for the "film" cameras. These units are arranged in two consoles facing each other and located close enough together so one operator has access to all camera controls. The consoles are open-ended with floor space and cable-rack capacity adequate to accommodate more than double the number of cameras (or other sources of video signals) initially installed. Equipment racks, in which are mounted amplifiers, video jack panels, and other units required for the distribution of telecine video signals, are located near the camera-control consoles. The patchcross unit (see below), whereby any projector-camera grouping can be assigned to any studio, is also located nearby.

1A. *Uniplex Projector-Camera Arrangement.* An innovation in the telecine projection area is the uniplex arrangement whereby a separate film camera is provided for each projector. Typical groupings of projector, camera, and rack of auxiliary equipment may be seen in Fig. 16. In a television studio plant the potential size of Television City, the uniplex arrangement makes it possible to handle peak load conditions with the lowest total investment in equipment. In the usual multiplex arrangement, in which two or more projectors are fed into each film camera, one or more of the projectors are always idle and lost as far as peak operations are concerned. The expense of this enforced idleness is seen when it is noted that a film camera costs less than half of the combined cost of the 16 mm and 35 mm projectors frequently grouped together in larger telecine installations. The idle-time cost differential is heightened by the relatively greater time out for maintenance which experience indicates is required for the cameras as compared to professional-grade projectors. The uniplex arrangement offers the advantage at all times of full flexibility in assignment, operation, and maintenance. In addition, it eliminates the quality deterioration sometimes attributable to optical multiplexing devices or to imperfect mechanical indexing arrangements used for multiplexing. Uniplex operation, furthermore, involves system provisions that are fully compatible with unitized reproduction devices such as flying-spot scanners and video-tape playback machines.

2A. *Patchcross Unit.* Assignment of a telecine projector-camera grouping to a studio entails the connection of audio, video, projector control, and communications circuits between the telecine equipments and the studio involved. Because of the many circuits involved, as well as the size of the ultimate Television City plant, it was found that employment of a conventional switching arrangement for this purpose would involve a system employing more than 70,000 switch or relay contacts.

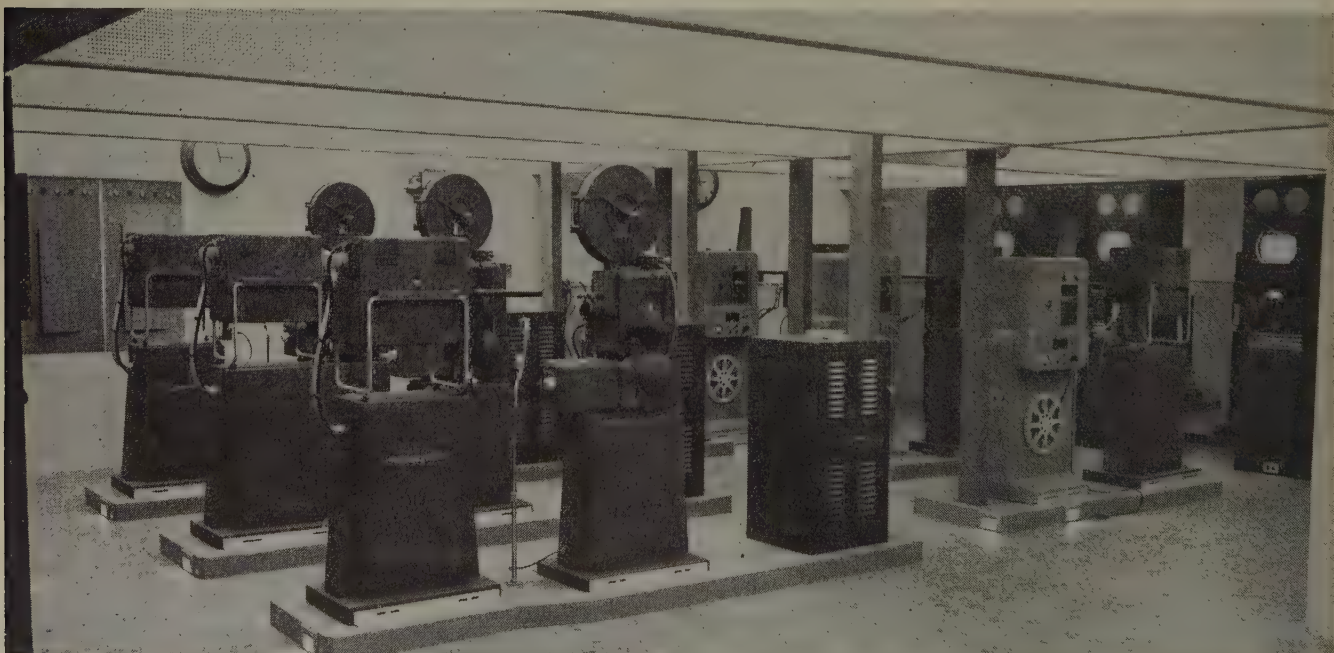

Fig. 16—An over-all view of the telecine projection room. The uniplex system, in which a separate camera is provided for each projector, can be seen. The acoustically treated ceiling is divided into cells over each equipment grouping to aid in localizing illumination and sound.

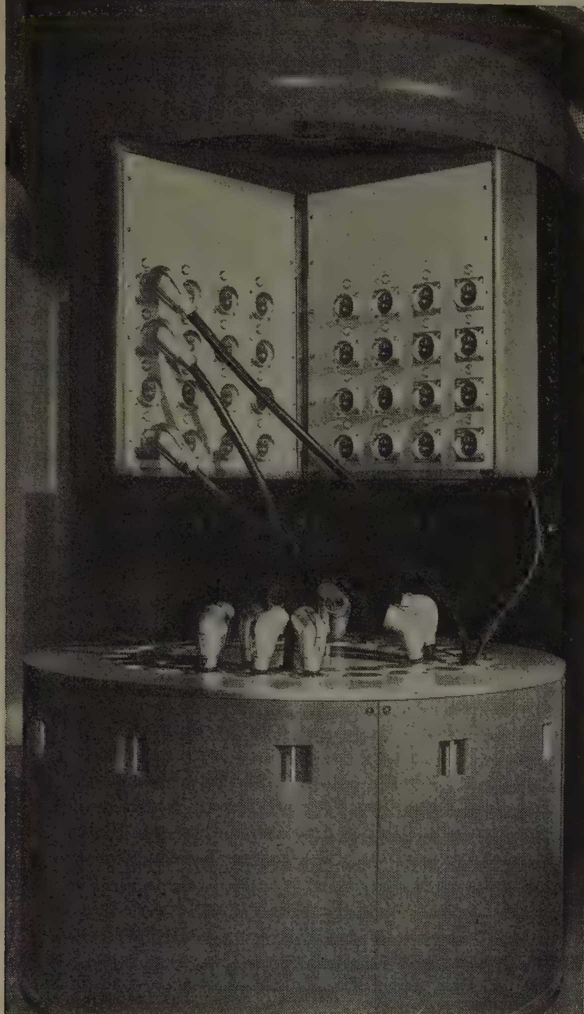

Fig. 17—The "patchcross" in the telecine control room serves to assign any telecine projector to any studio. All video, audio, communication, and projector remote control circuits are established in a single patching operation. Expansion space has been provided for plugs and receptacles to accommodate future projectors and studios.

For this reason, the circular patch panel (which is shown in Fig. 17) known as the "patchcross" unit, was developed. The patchcross unit employs multi-circuit connectors in a consolidated interconnection patching system wherein the audio, video, control, and communications circuits between a telecine projector-camera grouping and a studio are made in a single patchcord connection.

The lower section of the patchcross unit contains retractable patchcords, two being associated with each projector. The receptacles on the upper section are studio circuits, four such channels being provided to each studio and six to each program control room. Complete connection between any telecine projector station and any studio is made by patching the appropriate projector patch-plug into the desired studio receptacle. The second projector patchplug makes it possible to establish connections with a second studio on a preset basis while still connected with the first studio. The transfer from one studio to the other can then be made conveniently and quickly by means of transfer switches on the projector rack.

The unconventional form of the patchcross unit is a result of functional design considerations. The assembly was planned to be as small and compact as possible to keep the multi-conductor patchcords short and direct. This consideration resulted in adoption of the circular configuration. The "cross" shape of the upper section affords maximum panel space for mounting the 128 receptacles required to service the ultimate number of studios. The cylindrical lower section provides maximum usable volume for storage of retracted patchcords. The space between the upper and lower sections makes it possible for the operator to reach any projector patchplug from any point around the unit, and allows a short, direct patch from any patchplug to any receptacle.

B. Telecine Audio and Communication Facilities

The audio and communication facilities associated with each projector, together with a picture monitor, are mounted in the equipment rack which forms a component of the projector-camera grouping. Several of these projector racks are visible at the extreme right in Fig. 16. In these racks the various operating controls, as well as a volume indicator for visual monitoring of the film sound track, are centralized on a control panel directly below the picture monitor. Other audio and communication controls are located at the associated camera-control unit in adjoining telecine control room.

1B. *Telecine Audio Facilities.* The audio equipment associated with a typical film projector is shown in Fig. 18, page 1084. Sufficient gain has been provided, by preliminary and program amplifiers, to handle the audio level from film projectors in which no preamplification is provided. When these facilities are used with projectors having built-in preamplification, the plug-in preliminary amplifier is removed and a plug-in line transformer substituted in its place. This transformer permits the use of a balanced-to-ground audio circuit from projector to equipment rack which reduces to a minimum noise pickup from the high-voltage pulsed-light circuits used in some projectors.

The audio output of the program amplifier is divided two ways by a special differential network which has adequate attenuation to assure that each output is completely independent of the loading on the other output. One of the network outputs supplies audio to one of the two patchcross plugs; the other output supplies audio to the second patchcross plug.

The monitoring loudspeaker at the camera-control position in the telecine control room always monitors the audio output of the film projector. However, the monitoring loudspeaker at the film projector can be connected, by means of a monitor key switch, to monitor audio from the film projector or audio program material from either of the two studios to which it may be patched. Auxiliary contacts on the monitor key switch energize video relays to switch corresponding

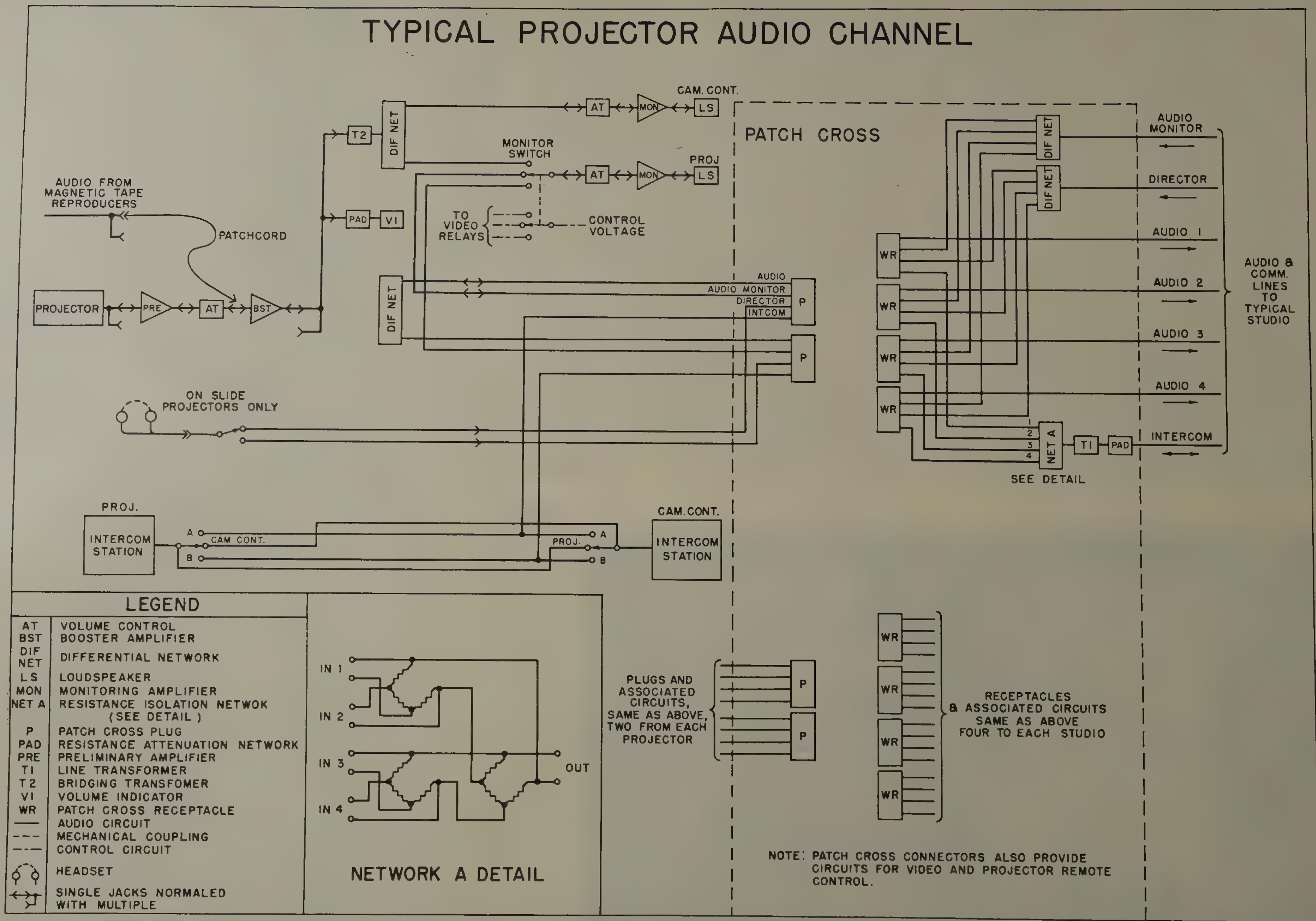

Fig. 18—This block diagram shows the audio and communication facilities associated with each telecine film projector. Connection of projector to any studio is made by the patchcross unit.

picture program material to the picture monitor on the projector rack.

2B. *Telecine Communication Facilities.* The communication facilities at a typical telecine projector are also shown on Fig. 18. Each station is equipped with a thermistor-controlled volume limiter. Circuit details of the individual intercom stations are basically the same as for the studio end of the circuit which has already been described.

As shown in the block diagram of Fig. 18, both the projector and camera-control stations are provided with a key switch by means of which they may talk to each other, or to either of the two studios to which the circuits may be connected through the patchcross system. It will be noted that the arrangement employed permits local communication between projector and camera control at all times except when each is switched to communicate with a different studio. Pilot lights at both projector and camera control indicate the setting of the switch at the other station.

During normal periods of operation two or more film projectors are usually patched to the same studio. Unless proper precautions are taken, under these circumstances the intercom conversation from one projector (or camera control) would be reproduced on the intercom loudspeaker at all other projector and camera-control stations patched to operate with the same studio. Inasmuch as these stations are located side by side, acoustic coupling between loudspeakers and microphone could result in "singing."

In the past, it has been customary to prevent singing from this cause by employing a relay interlocking system that silenced the loudspeakers of all such interconnected telecine intercom stations when one of the stations was transmitting. In the Television City installation a more direct and simple solution which requires no relays has been employed. Referring to Fig. 18, it will be noted that intercom lines from each projector associated with a given studio, after passing through the patchcross unit, are combined in a resistance isolation network, and the output of this network is transmitted to the studio. A detail of the resistance isolation network, which employs three Wheatstone bridge arrangements, is shown on the same illustration. A network of this type provides normal transmission (with a 12-db transmission loss) from any one of the four inputs to the output while providing a high degree of isolation between the four inputs. The isolation between the input circuits of each bridge element is a function of the accuracy of the bridge balance. It ranges from approximately 75 db with bridge legs having a tolerance of 0.1 per cent to approximately 30 db with bridge legs with 10 per cent tolerance. It is to be noted that the associated studio circuit forms a leg of one of the bridges.

In using a network of this type, care must be exercised that grounds on external circuits to which the network connects do not ground the network at points that would nullify its isolating characteristics. In this application, line transformers at each intercom station, as well as a line transformer at the output of the network, completely isolate the network from unwanted grounds.

It will be noted that the director's communication bus from each of the studios to which a projector is patched are brought directly to the audio jackfield on the projector rack by means of circuits through the patchcross unit. This communication bus is very useful when close co-operation is required between projectionist and studio and is almost always used in the operation of slide projectors. For this reason, in the case of the slide projectors, these circuits are extended directly to a selector switch and headphone jack on the projector.

Loudspeakers associated with the telecine emergency intercom circuits are mounted in each projector rack, in the rewind room, and at several locations in the telecine control room. When this emergency circuit from any studio is used, it can be easily and clearly heard in any part of the telecine area.

C. Telecine Video Facilities

Elements of the telecine video facilities appear in Fig. 19, page 1086. Contrasted with the audio system, in which each projector rack contains all amplifiers and distribution networks for the associated projector, the distribution of video signals involves a centralization of equipment. This is necessary because of the requirement that video-signal-circuit transmission time be uniform for all telecine camera channels in order that they may be superimposed in the various studio video switching units. The shortest possible over-all delay (for which compensation must be introduced in pulse circuits to all studios) is achieved by placing the camera controls, distribution racks, and patchcross unit in close proximity and keeping these elements reasonably close to the pulse distribution terminals in master control. The length of cable between film camera and film camera control does not enter into this problem, since the horizontal blanking pulse introduced in the camera-control unit covers the round trip delay of the narrower camera blanking pulse and becomes the timing reference signal to which studio camera-blanking signals must be matched at each studio.[9]

The close grouping of all video signal adjustment and patching elements, coupled with the reduction in operator manipulation of the camera-control units achieved by recent camera chain improvements,[10] make it possible for a small staff to cover the operation of the telecine control and projection rooms.

Video distribution of the film camera output signals is done on a matched impedance basis using two separately-powered video amplifier sections to feed the studios to which the camera is connected through the projector patchcords on the patchcross unit. The

[9] H. A. Chinn, "Television Broadcasting," McGraw-Hill Book Co., Inc., New York, N. Y., p. 46; 1953.

[10] K. B. Benson, "Feedback cascode iconoscope preamplifier," *Electronics*, vol. 26, no. 12, p. 166.

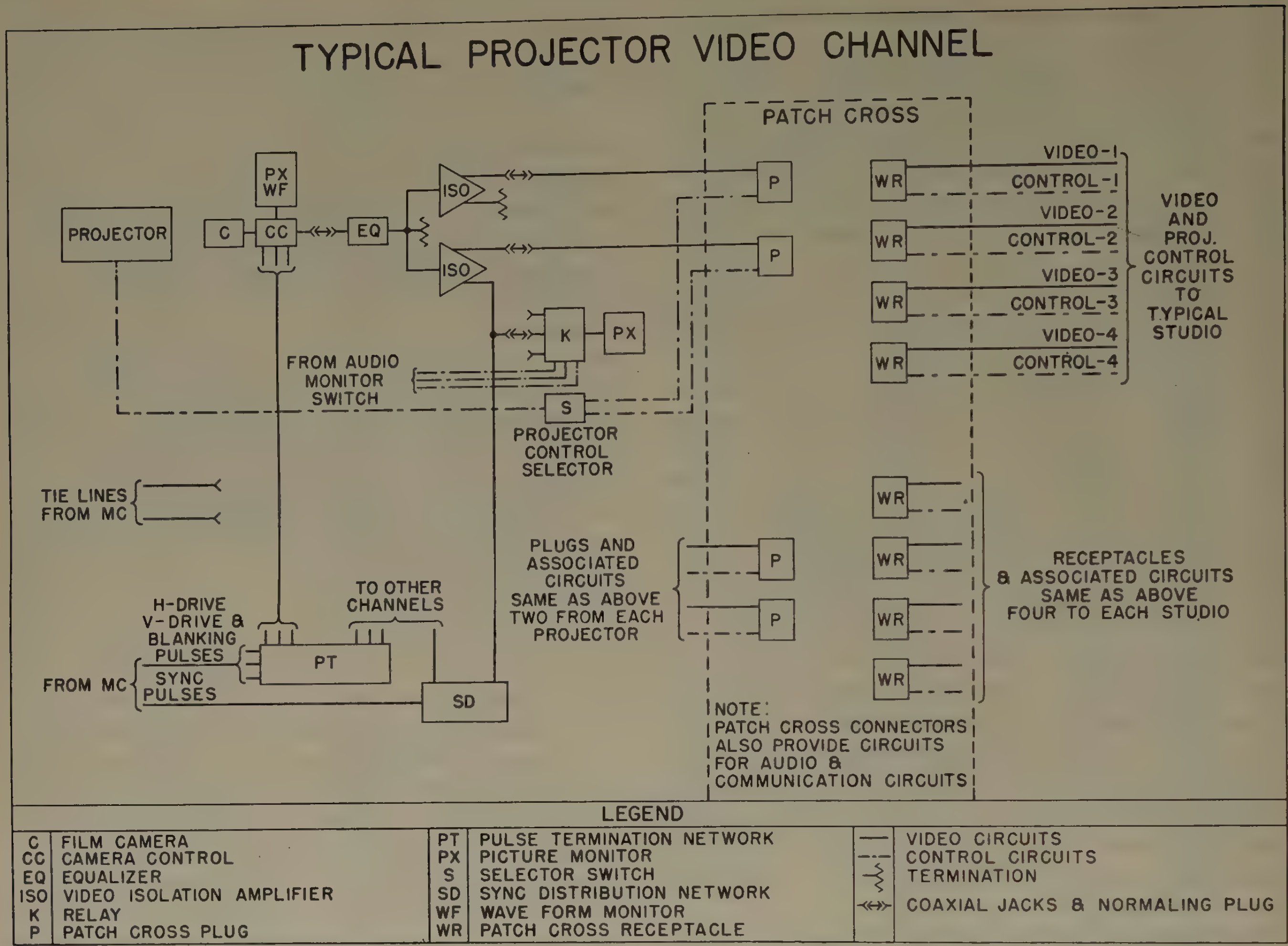

Fig. 19—Block diagram of the video facilities associated with a typical film projector. The patchcross unit (see Fig. 17) accommodates audio and communication circuits in addition to the video and projector remote control circuits shown above.

emergency coverage obtained by separate power supply connections for the two channels is augmented by spare amplifier groups which may be substituted for normally connected amplifiers by video circuit patching.

Equalizers are introduced between the camera-control unit and the amplifier inputs to pre-equalize for the coaxial cable losses on the median run to the existing studios. Future studios located farther away from telecine will be equipped with supplementary input channel equalizers and terminal amplifiers as needed.

A second output signal from one of the two amplifiers associated with each film camera is mixed with synchronizing pulses and fed to the picture monitor selector relay located in the projector rack. The projectionist may select for monitoring either the camera signal or studio cue.

The flexible assignment of projectors to the various studios made possible by the patchcross system requires that the remote-control circuits for all projector types utilize identical circuits. Unfortunately, at the present state of the art, the television industry does not enjoy any standardization whatsoever of television film projector control circuits, either internal or external. As a result, it was necessary to modify the control circuits of all projectors.

In modifying the Television City projectors to provide standardized control circuits, other features were incorporated, including the use of a mechanically latching relay which avoids the hazard of a lengthy hold-in circuit for projector operation, provision for local operation of the projector in the absence of control circuit power, and extension of single-frame still projection control to the studio to permit the continuous projection of a single frame of film.

Remote control of a projector may be patched to two control points at one time by means of the patchcross unit but only one of the two points can have control at any one time as assigned by the key switch on the projector rack. This prevents accidental miscontrol while permitting rapid reassignment of facilities.

As in the case of the studio installation, every effort was made in telecine to provide an adequate and reliable video system to handle present operating requirements and, at the same time, provide amply for future expansion. Since the centralized character of telecine requires a relatively greater measure of built-in expansion space, the equipment choice and layout were made to accommodate the estimated ultimate plant requirements. It is possible to expand readily the number of projector-camera groupings, to add synchronized mag-

netic-tape audio playback equipment, to install self-contained telecine scanners, and to add distribution connections to additional studios or additional circuits to present studios.

Master Control

The master control area of Television City (Fig. 20) has been designed to provide control, monitoring, switching, and terminal facilities capable of handling up to 50 incoming audio and video circuits from studios, remotes, and other program sources and distributing program material to as many as 12 outgoing program trunk circuits. In addition, a house monitoring selector system, also located in master control, will transmit audio and video signals to as many as 80 viewing rooms, offices, rehearsal halls, and other points. Also, in the master control area are the central timing-pulse generation and distribution system, the central terminal frames where all interconnecting audio, video, and control runs for the entire plant are terminated and cross-connected, and telephone company transmission terminal equipment.

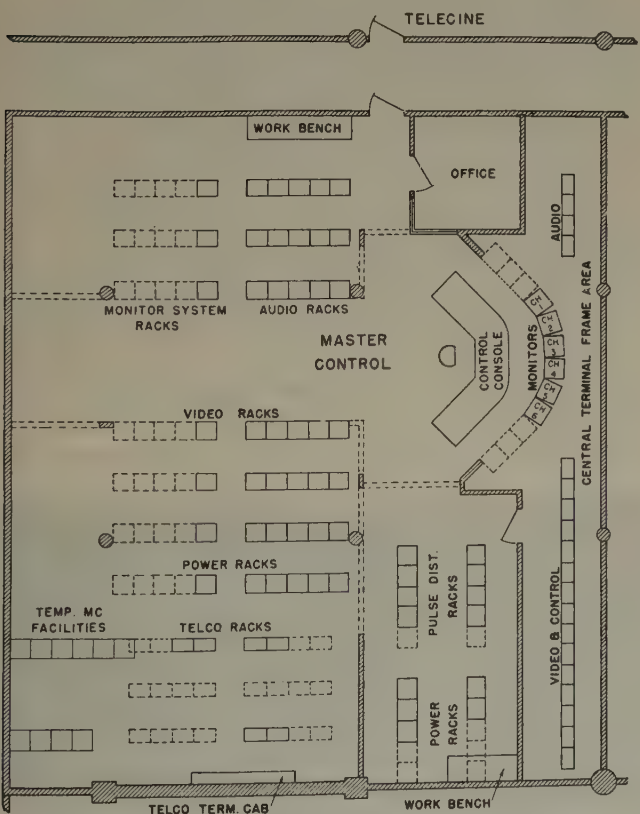

Fig. 20—Master control will accommodate control, switching, and monitoring facilities to handle up to 50 input channels and to feed as many as 12 output channels. The initial installation includes pulse generation and distribution facilities, telephone company terminal equipment, central terminal frames, and interim master control facilities.

In keeping with an over-all schedule for orderly transfer of operations from existing installations elsewhere in Hollywood, the master control switching facilities are to be installed at a later date. In the initial installation, the pulse distribution system, the central terminal frames, and the telephone company facilities were installed in permanent form. The initial audio and video terminal facilities, on the other hand, is an interim installation that provides the necessary equalization, level adjustment, cue signal distribution, jackfields, and utility equipment to service the initial group of studios. In the first phase of operation, the audio and video outputs from all four studios are normally fed through these facilities to an existing master control elsewhere in Hollywood, while cue and remote signals received from that point are distributed to studios and other points in Television City.

A. Pulse Distribution System

In a television studio plant, all cameras must be scanned in perfect synchronism to make possible the superimposition of picture signals from various sources and smooth transitions between signals. This necessitates the derivation of all timing signals from a central timing-pulse generation unit. Four basic pulses are transmitted from the master control pulse-distribution system in Television City to each studio: horizontal-rate driving pulses, vertical-rate driving pulses, blanking signals containing both horizontal- and vertical-rate components for mixing with the camera output signals, and composite synchronizing signals which contain the various horizontal- and vertical-rate pulses required in the final transmitted signal.

1A. *Timing Compensation.* A finite time, which is an appreciable fraction of the duration of a horizontal scanning line, is required for the transmission of pulse and video signals between points in a large studio plant. Because of this, it is necessary to introduce compensating delays to insure that signals from various points will be in phase at points where they are to be mixed or superimposed. In the Television City video system, delay is introduced at two points.

The first point is the horizontal-rate pulse feeds to each studio to bring the studio camera output signals into exact phase with telecine camera output signals which have been transmitted to that studio. The amount of delay time is equivalent to the time delay encountered by a blanking pulse, for example, in traveling from the pulse distribution point to telecine, through a camera control, and back to thec ommon terminal frame from which all runs to a studio depart.

This telecine-loop delay compensation is the same for all studios but is introduced in the circuits to each studio so that all pulse-distribution system outputs, including the one that normally feeds telecine, may be kept in phase at the origination point allowing full freedom in emergency patches.

The second delay compensation is made to bring the video output from all studios in phase at master control so the signals may be intermixed in a program control room, or in a remotely controlled video switcher in master control. This delay is accomplished by introducing lengths of cable in each studio output line, as re-

quired, to make the round-trip circuits from the pulse distribution racks, through the studios, and back to master control, identical for all studios. Delay is added for the physically nearest studios to provide a round-trip equal to that of the most distant studio. In the initial unit of Television City, this round-trip is equivalent to 1,100 feet of 75-ohm polyethylene-insulated coaxial cable.

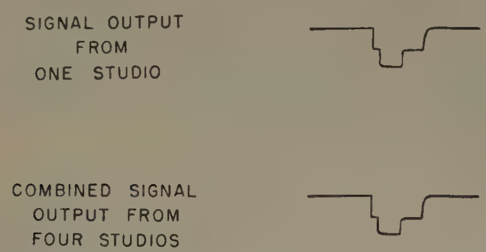

Fig. 21—These actual oscillograms show the effectiveness of timing compensation. Upper oscillogram shows a blanking signal from a telecine film camera routed through one studio to master control. Lower oscillogram shows the same signal routed to all four studios and received simultaneously at master control.

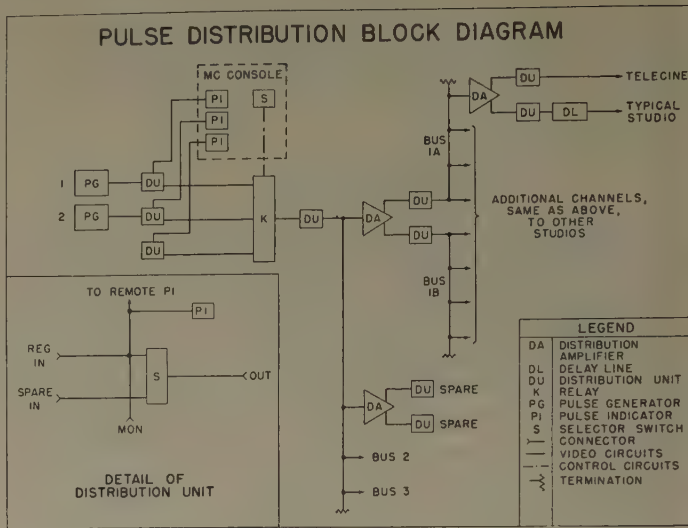

Fig. 22—Simplified block diagram of central pulse-distribution system. Although this simplified diagram shows the circuits for distribution of only one pulse signal, in reality there are four separate pulses distributed by the pulse distribution system, i.e., horizontal scanning drive pulses, vertical scanning drive pulses, blanking signals, and composite synchronizing signals.

The accuracy with which delays can be adjusted is illustrated in Fig. 21, which shows an oscilloscope display of a blank raster signal, obtained from a telecine camera, which was transmitted through studios, where synchronizing pulses were added, back to master control where the photograph was made. The combined outputs of four studios shows excellent registration.

2A. *Pulse Distribution.* The flexibility of the pulse distribution system is enhanced by application of an unusual patching and monitoring system shown in Fig. 22. The four-pulse circuits are grouped together throughout the system on multi-contact patching connectors. The four basic output pulses from either of the two separately-powered pulse generators may be selected for operation of the entire plant in an overlap-switching, mechanically-latching, video relay selector unit which may be operated from the master control console or other desired locations. The pulses from the selected generator are fed to a bank of main bus amplifiers, each amplifier chassis including channels to handle all four pulses. These, in turn, feed branch distribution amplifiers, the outputs of which feed the various studios.

At every important junction of the distribution system, the four pulse signals are routed through a special pulse distribution unit, a bank of which is shown in Fig. 23. Each distribution unit provides a monitoring receptacle for inspection of the pulse signals by means of a waveform or pulse-cross-display picture monitor, a switch for quickly substituting a set of previously patched spare input signals, and a signal-actuated neon-tube pulse-indicator unit. A neon tube, driven through a triode isolation amplifier from each pulse circuit, indi-

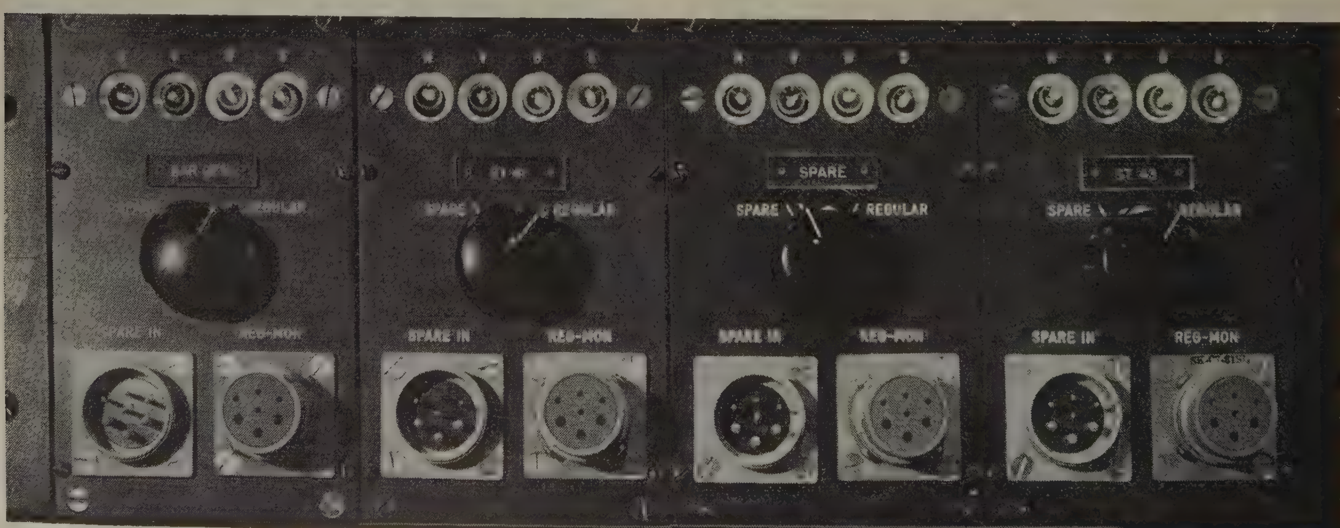

Fig. 23—Pulse distribution units, four of which are shown, are provided at each important junction point of the pulse-distribution system. They provide neon lamp visual monitoring indicators on each of the four pulse signals, access for oscilloscopic monitoring, and provisions for switching over to a pre-patched spare pulse feed in an emergency.

cates by a steady glow the presence of a signal or, by slowly blinking, the absence of a particular pulse. In this latter case, the triode isolation tube exhibits no ac output signal, and the neon tube blinks as it cyclically discharges the accumulating charge on a plate circuit capacitor. A third indication, in which all neon tubes are extinguished, occurs if the power unit which supplies both the indicator chassis and the associated pulse distribution amplifier should fail.

3A. *General.* The quickly interpretable indications of the pulse indicator and the ready means provided for obtaining alternate signal feeds are of considerable value in an extensive plant such as Television City because they greatly simplify operational trouble-shooting in this key central element of the video system. In keeping with the timing considerations noted above, and to help make possible unrestricted patching of pulse feeds, level adjustments are provided on each output of the distribution amplifiers.

A feature of the installation is the placement of the distribution amplifiers and the distribution unit monitoring facilities in facing rows of equipment racks making level adjustment an expeditious one-man operation. The equipment layout of this flexible pulse-distribution system, as in the case of telecine, has been arranged to accommodate the ultimate requirements of Television City, the components not required initially being simply omitted from the allocated spaces.

B. Other Features

Other features include the use of video equalizers for correcting the response-frequency characteristics of the studio video circuits, the use of a newly developed video terminal block, and of a flexible overhead cable supporting installation.

1B. *Video Line Equalization.* Even in the initial unit of Television City, the length of video cable runs required the use of equalization to achieve uniform response-frequency characteristics. The need for equalization is heightened by the introduction of additional cable lengths for delay purposes in the video circuits from studios to master control. Equalization was accomplished by following well-established transmission practice using passive-circuit attenuation equalizers in conjunction with flat-response video amplifiers to compensate for transmission losses. Equalization was handled in this manner, rather than by means of a special vacuum-tube peaking amplifier, in accordance with a policy of component standardization to reduce the number of types of amplifiers required.

Two types of equalizers, differing in the degree of high-frequency compensation, were designed for insertion in the one-side-grounded, 75-ohm video circuits that are employed on a matched impedance basis. Fig. 24 shows the circuit arrangement of the equalizers, together with the compensation characteristics obtainable from the two individually or in various combinations.

As applied in the interim master-control terminal

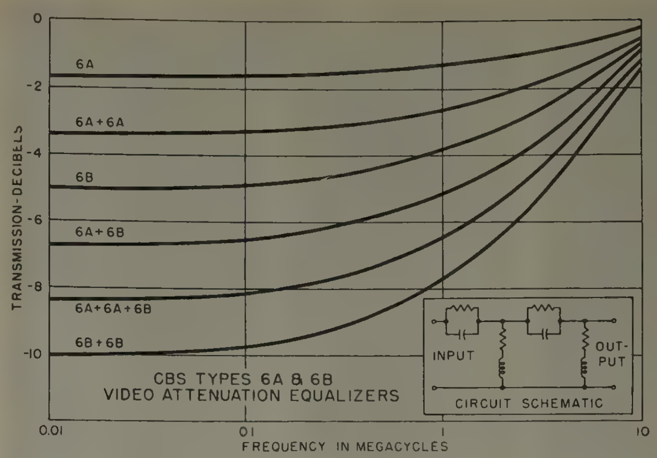

Fig. 24—Two attenuation equalizer types, differing only in degree of equalization, are used to correct the response-frequency characteristics of studio to master-control video circuits. The schematic circuit and compensation characteristics of the two equalizer types, as well as various combinations thereof, are shown.

facilities, the video program line from each studio is fed through an equalizer and an associated utility video amplifier which, in turn, supplies the telephone company facilities a properly equalized 1.4 volt, peak-to-peak, video signal. A second output from the same amplifier feeds a group of distribution amplifiers, the outputs of which may be connected to various monitoring points, as required, by means of patchcord connections. The second video program line from each studio is identically equalized and amplified. The entire second terminal channel may be substituted for the regular facilities, if desired.

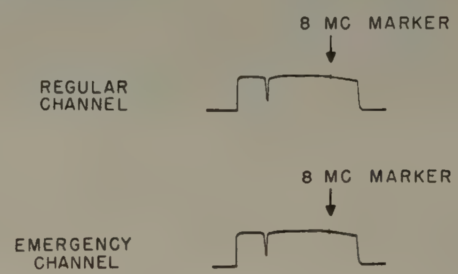

Fig. 25—Oscillogram showing typical response-frequency performance of a complete studio through master control.

The performance of a typical equalized studio circuit, including delay compensation, is shown in Fig. 25.

2B. *Video Terminal Blocks.* Coaxial cables at the central terminal frames, as well as throughout the entire Television City plant, have been terminated on solder-type video terminal blocks. This new terminal block, shown in Fig. 26, has been used in place of the coaxial connector junctions previously employed. These terminal blocks have proven to be faster and more economical to install and have the advantage of providing an accessible test point at all times. No measurable transmission discontinuities have been noted.

3B. *Cable Distribution.* As Television City expands, the addition of new studios or other technical facilities will require the installation of additional terminal equip-

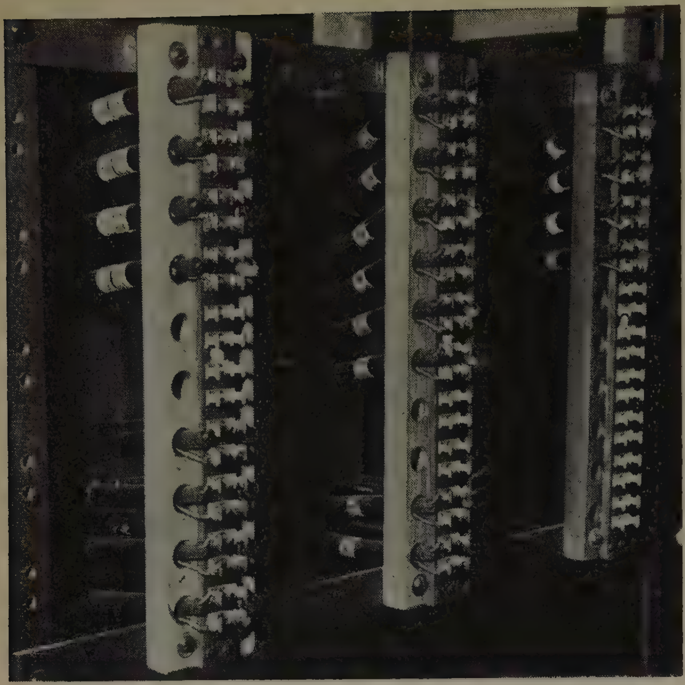

Fig. 26—Newly designed solder-type video circuit terminal blocks are used throughout Television City for the termination of coaxial cable circuits. No measurable transmission discontinuity in the video frequency range is introduced.

Fig. 27—The central terminal frames adjacent to master control. Space has been allocated on these frames for all video and communication circuits associated with future studios. A similar group of frames are provided for audio lines.

ment and interconnection wiring in the master control area. Such expansion will be greatly facilitated by the readily accessible overhead cable rack system and by the easily cross-connected audio, video, and control circuit terminal blocks on the central terminal frames adjacent to master control. Fig. 27 shows the section of the central terminal frames where video and control circuits are terminated. It also shows the overhead open ladder-type cable racks which carry the cable runs in this area.

Conclusion

The audio, video, and communication facilities for CBS Television City were designed in accordance with an over-all principle of providing a complete, but easily expansible, initial installation consistent with the basic flexibility inherent in the architectural design of the plant. This objective, together with the large physical size of Television City, led to an extensive decentralization of the video facilities, in keeping with well-established audio practice.

In addition to this major system innovation, a number of new technical practices and details have been introduced including: a tandem dual-fader studio video switching system, a flexible studio picture monitor bus-switching system, an effective installation of suspended audience picture monitors, a flexible unitized studio audio and communications system, a simple thermistor-controlled volume limiter on communication circuits, a uniplex film projector camera arrangement, a unique patchcross switching system for assignment of film projectors to studios, a standardized system of projector remote control, a flexible pulse-distribution system featuring signal-actuated pulse indicators, equalization of video lines, and a video terminal block.

In the application of these new ideas, as in the many adaptations of previously used techniques, emphasis has been upon sound and conservative engineering practice while, at the same time, providing adequately for future growth and expansion.

Acknowledgment

A project of this magnitude reflects contributions by many individuals. At the peak of activity, 60 persons were engaged in the audio and video engineering, drafting, construction, and installation work. The efforts of all persons concerned, including those associated with the architects, contractor, and suppliers, were marked by unusually effective co-operation. The authors wish particularly to express appreciation to W. B. Lodge, CBS Television vice-president in charge of engineering, A. B. Chamberlain, CBS Television chief engineer, L. H.

Bowman, CBS-Hollywood director of technical operations, and H. W. Pangborn, CBS-Hollywood manager of television operations, for their constructive advice and over-all guidance of the project; to K. S. Tyler, manager of CBS Television building construction for the very important team spirit which was engendered between construction and facilities groups, to our colleagues in the CBS Television Engineering Department, and to J. B. French, CBS-Hollywood manager of technical construction and maintenance, without whose vigorous on-the-job direction completion of the installation in record time would not have been possible.

A Developmental Medium-Power Traveling-Wave Tube for Relay Service in the 2,000-Megacycle Region*

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Summary—**This paper describes an experimental 2,000-megacycle traveling-wave power amplifier designed for operation at electrode voltages below 700 volts. This tube is capable of delivering a power output of 7.6 w at a gain of 20 db and a power output of 3 w at a gain of 28 db at the optimum operating point. Maximum power outputs from 4 to 7.6 w at gains ranging from 17 to 21 db have been measured over a bandwidth of 650 mc (1,700 to 2,350 mc). Electronic efficiencies of 20 per cent and collector efficiencies close to 30 per cent have been obtained with this amplifier. Results obtained with the tube operating as a frequency shifter are discussed. A design including a permanent magnet for electron-beam focusing has been developed.**

Introduction

THIS PAPER DESCRIBES a developmental 2,000-mc, medium-power, low-voltage, traveling-wave tube, illustrated in Figs. 1 and 2, designed primarily for use in the output stage of microwave relay transmitters. The development of this traveling-wave amplifier, started at the RCA Laboratories by Kaisel,[1] was undertaken in an effort to provide better performance in such applications than that obtainable from triodes and klystrons. This tube is also useful in frequency shifter or intermediate-stage-amplifier applications.

In the development of the traveling-wave amplifier, emphasis has been placed not only on electrical characteristics but also on achieving a satisfactory mechanical design. Efforts have been made to produce a tube which is shorter and more rugged than early experimental types. The following broad objectives have been taken into consideration:

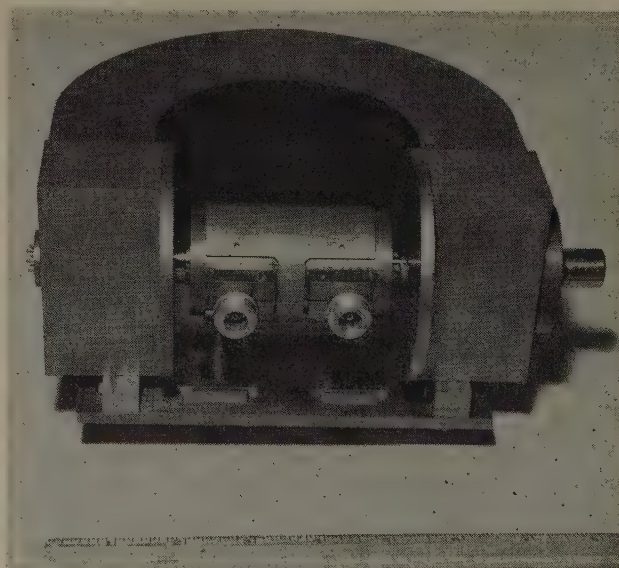

Fig. 2—View of traveling-wave power amplifier with permanent magnet and helix-to-coaxial-line transducers.

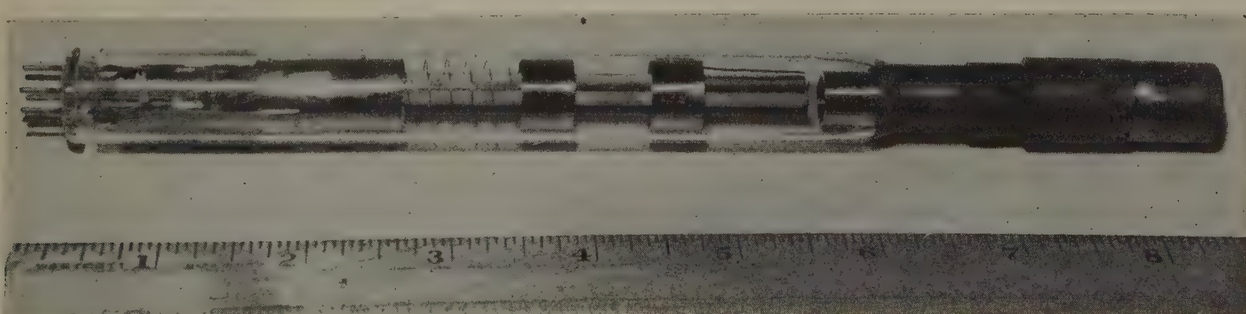

Fig. 1—Developmental 5-w 20-db traveling-wave tube for relay service in the 2,000-mc. region.

(a) Power output of at least 2.5 w.
(b) Gain of at least 20 db.
(c) Frequency range of 1,750 to 2,250 mc.

* Decimal classification: R339.2. Original manuscript received by the Institute, November 27, 1953.
† Tube Div., RCA, Harrison, N. J.
[1] W. J. Dodds, R. W. Peter and S. F. Kaisel, "New developments in traveling-wave tubes," *Electronics*; February, 1953.

(d) Operation at electrode voltages below 700 v.
(e) Coaxial input and output connections.
(f) A packaged design including a permanent magnet to focus the electron beam.

Items (d) and (f) of this list are of great importance in the application of the tube. Considerations of cost, reliability, and maintenance make it desirable to operate microwave relay links at low voltages. The use of a permanent magnet is desirable to eliminate bulky, power-consuming solenoids.

A helix-type circuit used in the traveling-wave amplifier satisfies the electrical specifications and lends itself, in this case, to a compact mechanical design. The voltage, frequency, and gain requirements are such that a helix about three inches long is necessary. With an air gap of this order, it is possible to use a conventional "U"-type magnet weighing from 10 to 15 pounds.

Severe requirements have been imposed on the electron gun. Because the efficiency of the tube is about 20 per cent, a high-current beam has to be generated at a low voltage. In addition, mechanical construction of the gun must be suitable for use with a permanent magnet.

Wide-band helix-to-coaxial-line transducers have been designed to feed energy into and out of the amplifier. The optimum dimensions and location of the attenuator used to prevent the tube from oscillating have been determined by experiment to prevent regeneration and to obtain maximum power, gain, and efficiency.

Tube Components

Helix Assembly

The helix, designed according to the theory described by J. R. Pierce,[2,3] consists of 0.010-inch tungsten wire wound at a pitch of 0.0184 inch; its inner diameter is 0.120 inch. With this helix and the beam produced by the electron gun which will be described later, the gain parameter, C, referred to by Pierce is about 0.15 at a collector current of 50 milliamperes. The main features of the tube assembly are shown in Fig. 3. Three ceramic rods used as supports for the helix are held in position by clamps welded to the matching cylinders. The antennas and the matching cylinders are components of the input and output coupling circuits. The last two turns at both ends of the helix are stretched to provide a smoother transition to the straight antenna and thus to improve the match.

The electron gun and the helix assembly comprise a unit structure centered with respect to the collector. The collector fits into the aligning elements located in the pole pieces, as shown in Fig. 3. In this manner, the electron gun and the helix are in the right position for beam focusing when the tube is inserted into the magnet. The structure is rugged and easy to construct, and permits a high pumping rate during the exhaust operation. Dielectric loading is kept to a minimum, and attenuation can be conveniently applied directly to the helix.

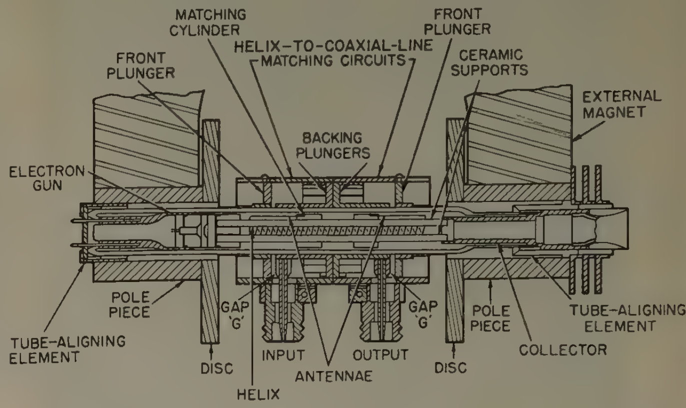

Fig. 3—Schematic diagram of traveling-wave power amplifier assembly.

Electron Gun

The magnetically confined, "parallel-flow" electron gun used in this tube is mechanically simple and satisfies the electrical requirements. Because the gun structure does not provide beam convergence, the mean current density at the cathode is practically the same as that of the beam. With this design, it is possible to obtain a high perveance [collector current/(positive electrode voltage)$^{3/2}$] and to make the components of the gun small. A perveance of approximately 5×10^{-6} amperes per (volt)$^{3/2}$ is necessary because of the comparatively low voltage requirement and the fact that the electronic efficiency of the tube is about 20 per cent.

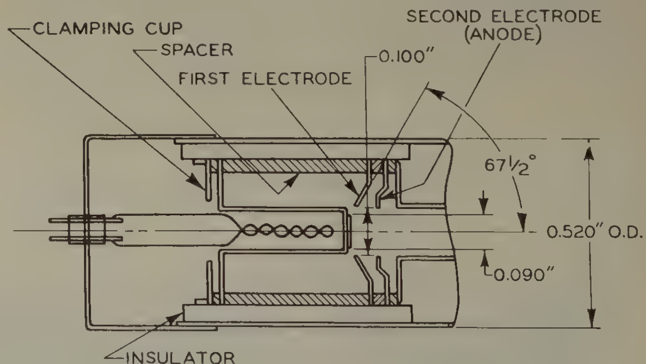

Fig. 4—Diagram of the electron gun.

The electron gun, shown in Fig. 4, has been designed according to the method described by J. R. Pierce.[4] The cathode used in this gun has a diameter of 0.090 inch. Two electrodes are used to focus the electron beam. The first electrode is usually maintained at the same potential as the cathode. The second electrode, or anode, is positive with respect to the cathode. In the development, it was assumed that the space-charge conditions are the same as those in a parallel-plate diode. The

[2] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y., 1950.

[3] J. R. Pierce, "Theory of the beam-type traveling-wave tube," Proc. I.R.E., vol. 35, pp. 111–123, February, 1947.

[4] J. R. Pierce, "Theory and Design of Electron Tubes," D. Van Nostrand Co., Inc., New York, N. Y., 1949.

electrodes are shaped to cause the potential along the beam boundary to vary as the 4/3 power of the distance from the cathode. This type of potential distribution maintains the boundaries of the electron-beam parallel.

In this design, the cathode of the electron gun is usually in the magnetic field. Even in the absence of a magnetic field, however, the current intercepted by the electrodes is negligible provided the potential of the helix is higher than that of the second electrode.

The mechanical construction of the electron gun permits the use of small radial dimensions. This feature is desirable because the tube has to fit inside the magnet pole pieces. The diameter of the holes in the pole pieces should be as small as possible so that the necessary field strength and uniformity can be obtained.

Helix-to-Coaxial-Line Transducers

The construction of the helix-to-coaxial-line transducers is shown in Fig. 3 and in the upper right-hand corner of Fig. 6. In the development of these transducers, it was desired that they should not add to the axial length of the tube. This objective was achieved by "folding back" such components as the antenna and the matching cylinders over the ends of the helix, as shown in Fig. 6. With this design, other elements, such as the electron gun and the collector, can be placed close to the helix assembly without absorbing energy from it. Because the spacing between the collector and the output end of the helix is small, it is possible to use a collector voltage lower than that of the helix without reducing the power output. As a result, collector efficiencies [(rf power output × 100)/(collector voltage × collector current)] of 30 per cent have been obtained.

The rf power is fed into the helix from the antenna, which is energized through the gap "*G*," as shown in Fig. 3. The desired impedance match is obtained by adjustment of the front plunger to provide the proper width of the gap "*G*" and by suitable location of the antenna with respect to the gap.

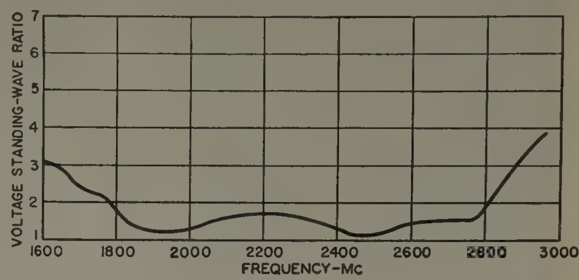

Fig. 5—Characteristic of voltage standing-wave ratio vs frequency for helix-to-coaxial-line transducers.

Fig. 5 shows a curve of the voltage standing-wave ratio with respect to frequency for the helix-to-coaxial-line transducers. In these measurements, an attempt was made to obtain a voltage standing-wave ratio of less than 2 over as wide a frequency range as possible. Standing-wave ratios of less than 1.8 were measured over a bandwidth of 1,000 mc, and ratios below 3 over a bandwidth of 1,300 mc.

In many applications, very low standing-wave ratios are required over a narrow bandwidth. Tests have been made to determine whether the type of helix-to-coaxial-line transducers described above is suitable for use in such applications. The results of these tests are shown in Fig. 6. For the curves at the bottom of Fig. 6, an attempt was made to obtain standing-wave ratios of about

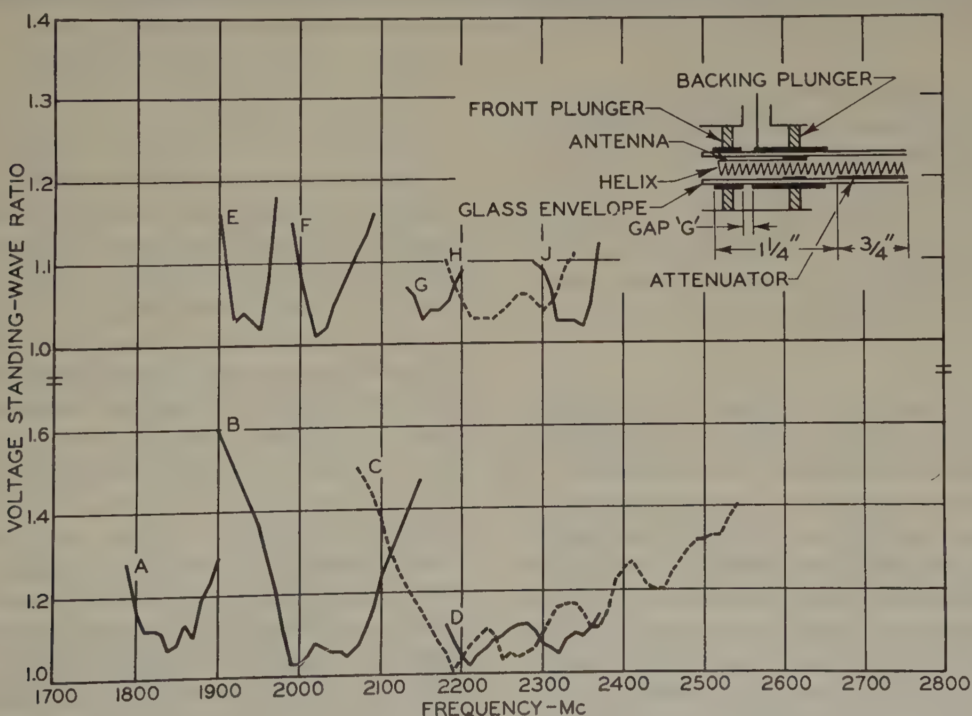

Fig. 6—Characteristic of voltage standing-wave ratio vs frequency for helix-to-coaxial-line transducers. This figure shows the bandwidths over which standing-wave ratios of approximately 1.1 or better can be obtained.

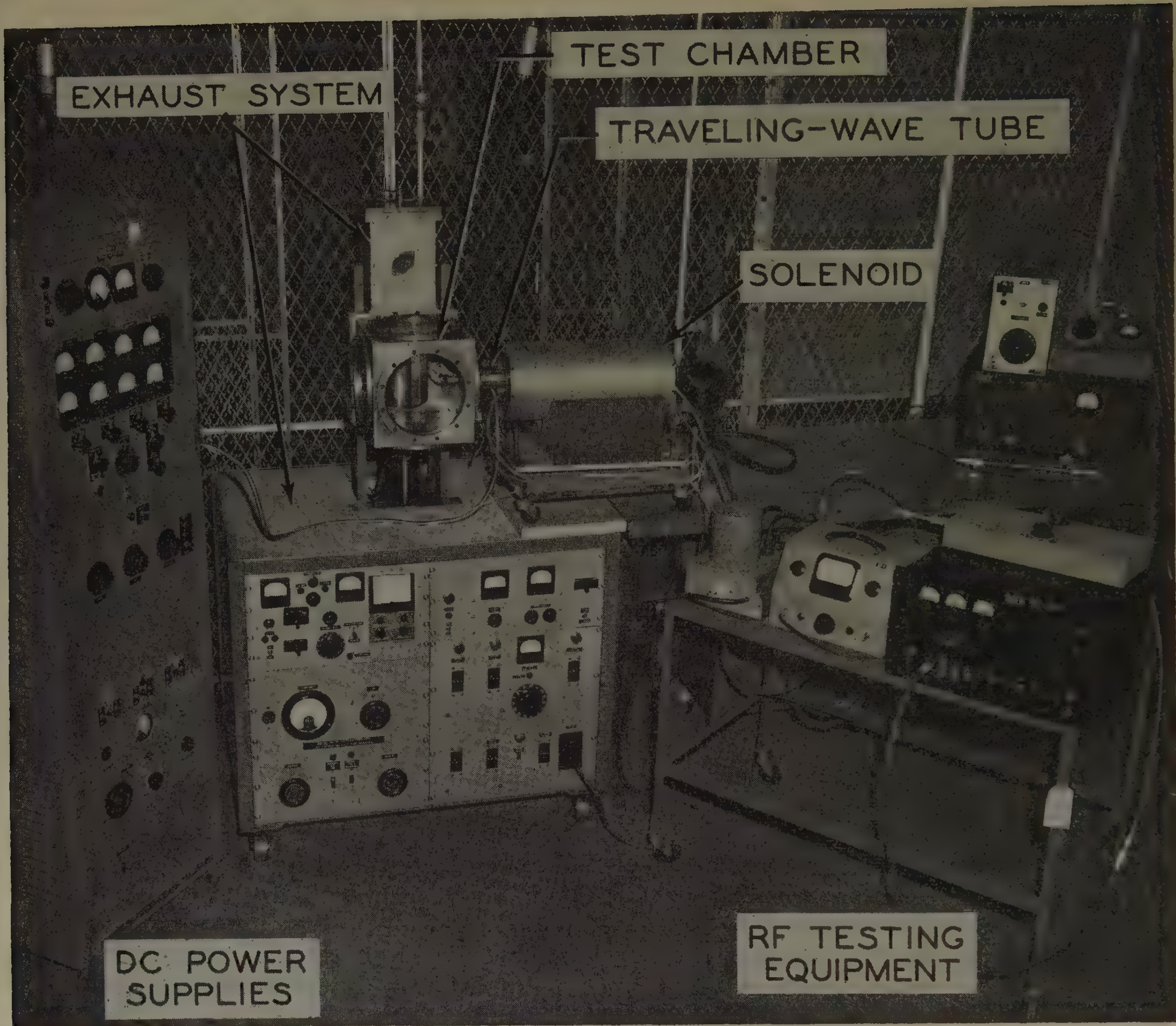

Fig. 7—Demountable exhaust system and testing equipment.

1.1 over as wide a frequency band as possible. Standing-wave ratios of less than 1.12 were obtained over bandwidths of 100 and 185 mc for curves "*B*" and "*D*," respectively.

The curves at the top of Fig. 6 show the results of tests to obtain standing-wave ratios of less than 1.1. It can be seen that standing-wave ratios of less than 1.07 have been realized over bandwidths of at least 50 mc.

Attenuator

Because most traveling-wave tubes of the helix type show a strong tendency to oscillate, an attenuator is used to prevent this condition. For good results, the attenuator should introduce very small reflections, be capable of dissipating the power, and be thoroughly degassed. Aquadag,[5] deposited on a ceramic rod, is used to provide the required loss characteristic. A "cold" insertion loss of about 40 db is necessary for the traveling-wave amplifier described in this paper. The dimensions and the location of the attenuator, which were determined by experiment, are shown in the upper right-hand corner of Fig. 8.

[5] Trademark registered by Acheson Colloids Co., Port Huron, Michigan.

Experimental Results

Experimental Techniques

Preliminary evaluation of circuits was carried out in a continuously pumped demountable system,[6] shown in the center of Fig. 7. The use of the demountable system effects a considerable reduction in the amount of time required for tube development. The tube and associated circuits are mounted on plates and sealed in the openings in the sides of the cubical test chamber. In case of failure of the tube under test, the test chamber can be isolated from the diffusion pump by means of a valve, and the assembly can be removed quickly and easily for repairs. The repaired assembly, or a new assembly, is then mounted in the test chamber, and the chamber is exhausted rapidly by means of a mechanical pump. The valve separating the test chamber and the diffusion pump is opened and an adequate vacuum for tube operation is obtained within a few minutes from the time of mounting the new assembly. In this manner, a large number of structures can be evaluated in a much shorter

[6] T. M. Shrader, "A demountable vacuum system for electron-tube development," Proc. of the National Conference on Tube Techniques sponsored by the Panel on Electron Tubes, October, 1953.

time than would be possible if each tube had to be exhausted and sealed before test. A solenoid was used in the demountable system to produce the magnetic field for focusing the electron beam. Results shown in Figs. 8 through 13 were obtained in the demountable system.

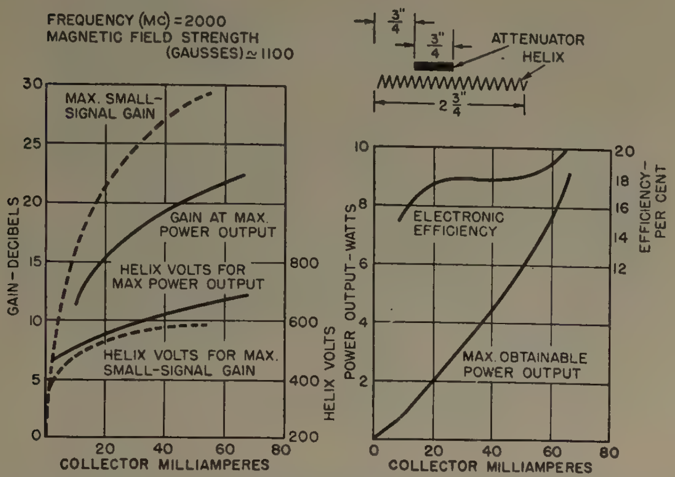

Fig. 8—Gain, power output, efficiency, and optimum helix voltages vs collector current. Solid lines indicate large-signal characteristics; dashed lines indicate small-signal characteristics.

Tube Characteristics

Fig. 8 shows the power output, the gain, and the optimum helix voltages as functions of the collector current for the traveling-wave amplifier. These measurements were taken at a frequency of 2,000 mc. The maximum power output shown in Fig. 8(b) can be obtained at a given collector current by optimum adjustment of the driving power and the helix voltage. The tube can deliver a power of several watts at a gain of about 20 db, depending on the current. The maximum obtainable power, P, can be expressed empirically as follows:

$$P = 0.037 \text{ (collector current in milliamperes)}^{1.31}.$$

The electronic efficiency of the tube given in Fig. 8(b) [(rf power×100)/collector current×helix voltage)] is approximately 18 per cent at collector currents ranging from about 20 to 50 ma, and increases to 20 per cent at 65 ma.

The maximum small-signal gain at 54 ma is 29 db, as shown by the upper dashed curve in Fig. 8(a). The helix voltage for maximum low-level gain is less than the voltage for maximum obtainable power output. Operation below 700 v has been achieved at both levels of operation.

The rf power output of the traveling-wave amplifier is shown as a function of the rf power input in Fig. 9. The bottom curve, A, in this figure was obtained with helix voltage and the driving power adjusted for maximum obtainable power output at a constant collector current. A maximum power output of 7.6 w was obtained at a gain of 20 db. The gain at the helix voltage corresponding to the maximum obtainable power output was found to increase as the input power was reduced, reaching 24 db at small signals and remaining practically constant for power outputs to about 4 w. Operation of the tube under the conditions used for these curves is suitable in applications where sufficient driving power is available and a high power output, achieved at some loss in gain, is desired.

When a high gain is preferable to maximum power output, operation before the saturation point is recommended. The dashed curves in Fig. 9 were obtained by reducing the driving power and adjusting the helix voltage, shown in the center curve, for maximum power output. It can be seen that the tube delivered 5 w of power at a gain of 26 db, 3w at 28 db, and 1w at 29 db.

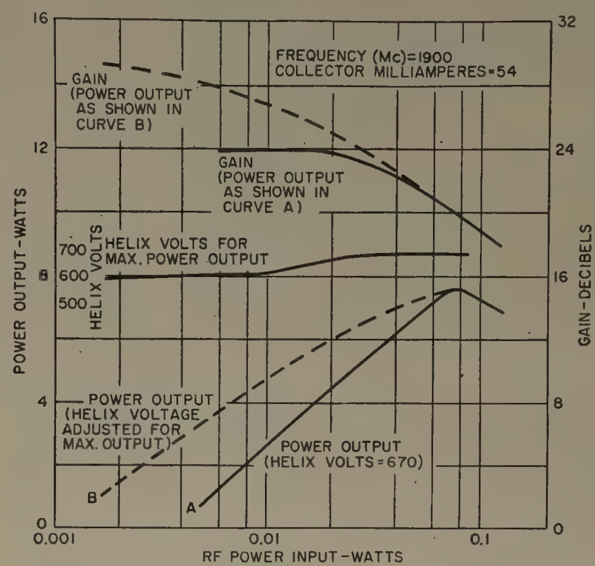

Fig. 9—Rf power output and gain vs rf power input. Helix dimensions and attenuator position as in Fig. 8.

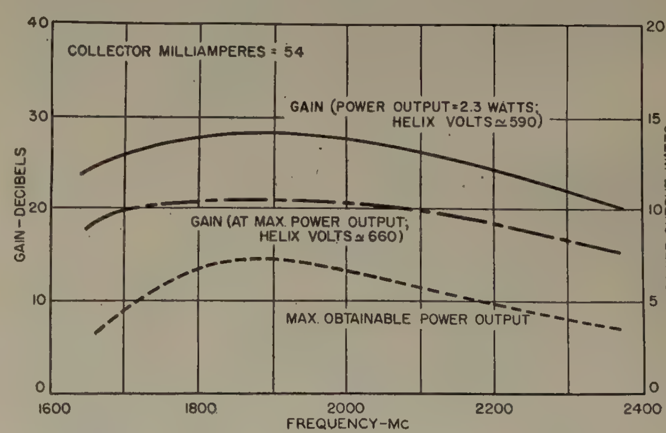

Fig. 10—Rf power output and gain as functions of frequency. Helix dimensions and attenuator position as in Fig. 8.

Fig. 10 shows the gain and power output of the tube as functions of frequency. The top curve in Fig. 10 shows the gain when the power output is adjusted to 2.3 w at each frequency. The maximum gain in the center of the frequency band was 28 db; the gain dropped to 21 db at 2,360 mc and to 25 db at 1,670 mc.

The maximum obtainable power output and the corresponding gain are shown in the lower curves in Fig. 10. Maximum obtainable power outputs ranging from 4 to 7.6 w at gains from 17 to 21 db were measured over a

frequency band from 1,700 to 2,350 mc. These results were obtained at a constant collector current.

A collector efficiency higher than the electronic efficiency can be obtained by the use of a collector voltage lower than that of the helix. The use of as low a collector voltage as possible is desirable to keep the power dissipated on the collector low. Fig. 11 shows the variation in power output and collector efficiency with collector

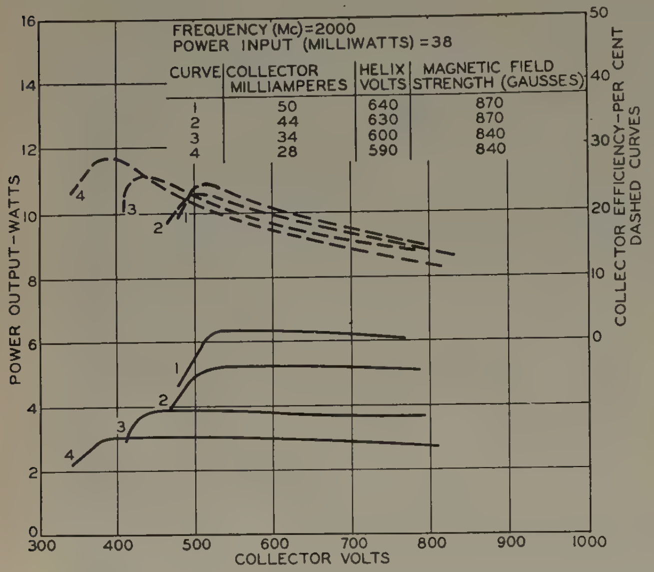

Fig. 11—Rf power output and collector efficiency as functions of collector voltage. Helix dimensions and attenuator position as in Fig. 8.

voltage. The data for the curves of Fig. 11 were obtained with constant collector currents, magnetic field strengths, and helix voltages. At each value of collector current, the helix voltage and the driving power were adjusted for maximum power output with the collector about 200 v positive with respect to the helix. The

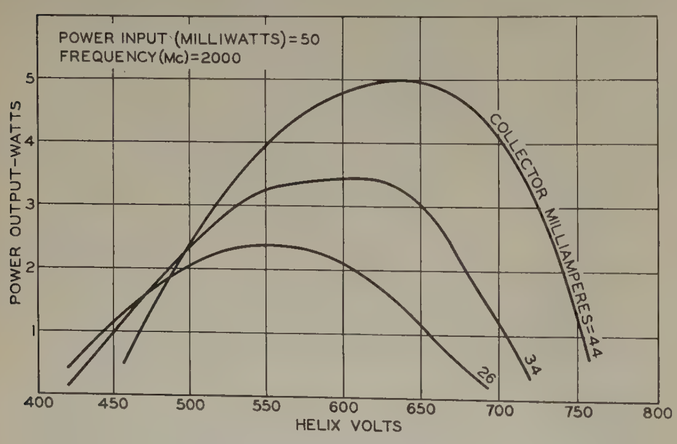

Fig. 12—Rf power output as function of helix voltage at constant collector currents.

curves indicate that the power output remains essentially constant until some minimum value of collector voltage, depending on the collector current, is reached; at that point, a rapid drop in power output is noticeable. Curves 1 and 3 in Fig. 11 show that at collector currents of 50 and 34 ma reductions in collector voltage of 120 and 160 v, respectively, do not reduce the power output. At power-output levels from 3 to 6.5 w, maximum collector efficiencies from 23 to 29 per cent were realized. These results were obtained at magnetic field strengths from 840 to 870 gausses.

Fig. 12 shows the power output as a function of the helix voltage for three values of collector current. It can be seen that the "3-db voltage bandwidth" at the current of 44 ma was 230 v.

Results shown in Figs. 8 through 12 were obtained by maintaining the first electrode at the cathode potential and by drawing the current with the positive potential on the second electrode.

An appreciable improvement in gain and efficiency can be achieved by an increase in the current density in the region where the axial component of the rf electric

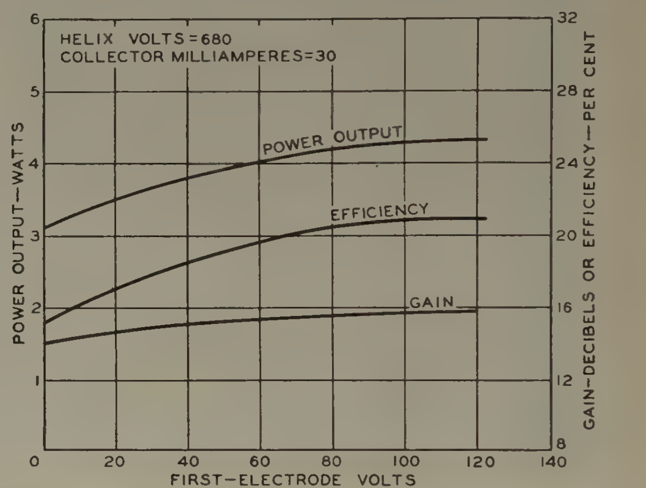

Fig. 13—Rf power output, gain, efficiency, and helix voltage as functions of first-electrode voltage. Collector current held constant at 30 ma by adjusting the second-electrode (anode) voltage. Helix dimensions and attenuator position as in Fig. 8. This test was performed on a different assembly than the one used to obtain results shown in Figs. 8 through 12.

field is of high amplitude. Because the field inside the helix increases with the radius,[2,4] the interaction between the electron beam and the field can be improved by the application of a positive voltage to the first electrode to produce a current-density distribution along the cross section of the electron beam which also increases with the radius. The results obtained by operation of the electron gun in this manner are shown in Fig. 13. With a collector current of 30 ma and the first electrode at cathode potential, the output was adjusted initially to 3.1 w. The curve of power output vs the first-electrode voltage, with the collector current maintained at a constant value by adjustment of the second-electrode (anode) voltage, shows that the power increases with the first-electrode voltage. The output was 40 per cent higher with a first-electrode voltage of 110 v than with voltage of zero.

A summary of the performance data obtained in the demountable system in Fig. 7 is given in Table I.

TABLE I

Voltages	below 700 v
Collector Current	54 ma
Magnetic Field	800 to 900 gausses
Max. Obtainable Power Output	4 to 7.6 w at 17 to 21 db gain measured from 1,700 to 2,350 mc
At Optimum Operating Point:	
Max. Power Output	7.6 w at 20 db gain at 1,900 mc
Gain at Power Output of 3 w	28 db
Small-Signal Gain	29 db

Frequency-Shifter Application

The theoretical considerations and actual operation of the traveling-wave amplifier as a frequency shifter have been described by W. J. Bray.[7] A suitable circuit for this type of operation is shown in Fig. 14. Results obtained with a sealed-off tube operating in this circuit are shown in Fig. 15. In this type of application, the cathode and the focusing electrodes of the electron gun are effectively grounded at the modulating frequency of 70 mc and the modulating voltage is applied to the helix.

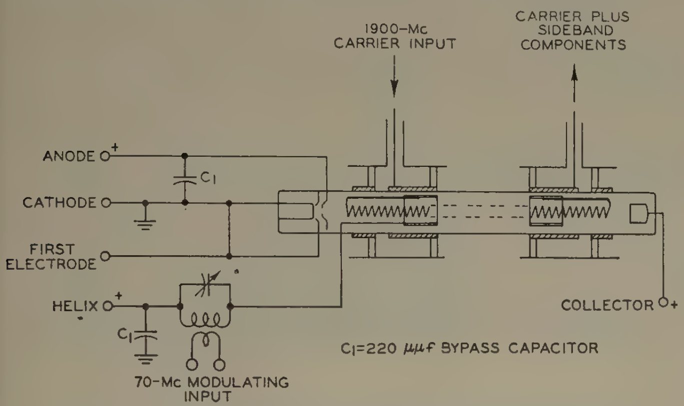

Fig. 14—Circuit diagram for operating the traveling-wave amplifier as a frequency shifter.

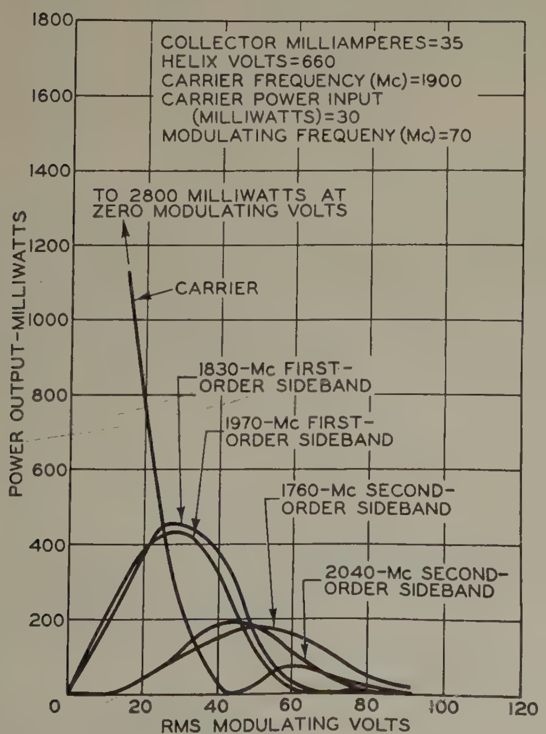

Fig. 15—Power output of carrier and sideband components as functions of sinusoidal modulating voltage.

[7] W. J. Bray, "The traveling wave value as a microwave phase-modulator and frequency-shifter," *Proc. Inst. Elec. Eng.*, vol. 99, part 3; January, 1952.

The operating conditions used to obtain the results shown in Fig. 15 are given at the top of the figure. It can be seen that the first-order lower sidebands reached a maximum at the modulating voltage of about 27 v (rms). The conversion gain (sideband power output/carrier power input) for the first-order sidebands is approximately 11.7 as compared to a gain of 19.6 db when the tube is operated as an amplifier. The first-order lower sideband is thus 7.9 db lower than the amplitude of the unmodulated carrier. Complete suppression of carrier occurred at modulating voltage of 44v (rms).

"Packaged" Design

An important feature of the traveling-wave amplifier, shown in Fig. 1, is the uniform diameter of the glass envelope. This design was possible because of the small diameter of the electron gun and the "folded back" design of the helix-to-coaxial-line transducers. This construction adds greatly to the mechanical strength of the tube, and eliminates the use of long and fragile glass tubing such as that used to support the helix in earlier traveling-wave tubes. Outside diameter of the envelope is 0.700 inch, and total length of tube $8\frac{1}{2}$ inches.

The "packaged" design is shown in Figs. 2 and 3. The supports for the matching circuits are omitted in these figures for the sake of clarity. The tube is self-aligning in the permanent magnet, as described previously in connection with the helix assembly. The magnet illustrated weighs 13.5 pounds and produces a field of 850 gausses on the axis of the tube. Circular discs are attached to the pole pieces of the magnet to improve the uniformity of the field.

The performance of sealed-off tubes has been similar to that obtained in the demountable vacuum system. Results with the permanent magnet, however, have thus far been somewhat inferior to those realized with the solenoid. The power output has been about 30 per cent less and the gain approximately 2 db lower. The difference in performance is believed to be due to the fact that the field in the permanent magnet is not as uniform as that in the solenoid. It is expected that suitable shaping of the pole pieces and discs will make it possible to obtain a more uniform field.

Acknowledgment

The author wishes to give credit to the RCA Laboratories in Princeton, N. J., and in particular to S. F. Kaisel, former member of the RCA Laboratories, for original development work on the traveling-wave amplifier, to the Advanced Development Group of the RCA Engineering Products Division in Camden, N. J., for evaluation of the tube as a frequency shifter, and to B. B. Brown, H. K. Jenny, and other members of the RCA Tube Department, for their contributions to the development of the tube described in this paper.

IRE Standards on Television: Methods of Measurement of Aspect Ratio and Geometric Distortion, 1954*

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Part I

1. Introduction

In a television system, it is through the agency of the scanning process that the two-dimensional space function, comprising the image at the camera, is transformed into a one-dimensional time function for transmission; at the receiver, the reverse procedure is used to recreate the image. Ideally, the velocities of the scanning apertures should be uniform in both the horizontal and vertical directions; furthermore, these two directions of motion should be orthogonal at all points in the raster. Finally, the ratio of the maximum excursions of the apertures in the horizontal and vertical directions of the transmitted picture, i.e., the aspect ratio, should be the same at both the transmitter and receiver. If there is departure from any of these conditions, geometric distortion results (except in the trivial case where the transmitter and receiver happen to contain compensating errors). It should also be noted that even in the case where the scanning aperture itself meets the desired conditions, the optical systems may contribute to the total geometric distortion; this applies to either the transmitter or receiver.

1.1 General Description

Measurement of geometric distortion involves comparison of an electrically generated "time pattern" with an optically or mechanically produced "space pattern." These can be of various types, and either fixed or movable, depending upon the purpose and type of the measurement. The method of measuring the image-pickup device will necessarily be slightly different from that used in the picture-reproducing monitor, since only in

* Reprints of this Standard 54IRE23.S1, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y. at $0.60 per copy. A 20-per cent discount will be allowed for 100 or more copies mailed to one address.

the latter case can the actual trace of the scanning beam be seen. For the camera, suitable charts or slides must be used, which produce the necessary optical pattern. For checking the monitor, the time pattern can be related to the space pattern by direct measurement of the former as its appears on the screen. Such charts and space patterns are commonly called "Ball" charts.

It is the purpose of this Standard to describe methods whereby the aspect ratio of the camera or other image-pickup device as well as that of the picture monitor may be properly adjusted. Methods will be described whereby, for a camera or monitor using the present U.S. standards for monochrome television, it may be determined if the geometric distortion at any point in the picture, whether electrical or optical in origin, exceeds certain numerical tolerances. Furthermore, within limits, the numerical amount of distortion may be measured. The procedures described in this Standard must necessarily be of a somewhat general nature, since the specifications of various equipments may vary somewhat. The application to a particular design of apparatus should be obvious, however.

Further methods of measurement of geometric distortion, of a more advanced nature, will appear in later Standards.

For the convenience of the user, this Standard is divided into two parts: the first describes methods for checking the picture monitor; the second concerns the image-pickup device.

1.2 *Definitions*

Geometric Distortion: Any aberration which causes the reproduced picture to be geometrically dissimilar to the perspective plane projection of the original scene.

Aspect Ratio: The ratio of the frame width to the frame height.

Scanning: In television, the process of analyzing or synthesizing successively, according to a predetermined method, the light values of picture elements constituting a picture area.

Scanning Linearity: The uniformity of scanning speed during the trace interval.

Raster: A predetermined pattern of scanning lines which provides substantially uniform coverage of an area.

Time Pattern: A picture-tube presentation of horizontal and vertical lines or dot rows generated by two stable frequency sources operating at multiples of the line and field frequencies.

Space Pattern: A geometrical pattern appearing on a test chart designed for the measurement of geometric distortion.

2. Measurement of Picture Monitor

2.1 *Apparatus and Circuits*

2.1.1 Basic Methods: The most convenient method of measuring geometric distortion is to measure the position of the aperture at regular intervals of time. These intervals can best be marked by means of suitable periodic electrical signals. In particular, two sets of intervals are used, each representing some fraction of the horizontal and vertical deflection periods, respectively. When this signal is applied to the monitor in place of the usual picture signal, a regular pattern of lines or dots appears on the picture tube, which is referred to as a time pattern. If there is no geometric distortion in the monitor, this time pattern will consist of straight lines or rows of dots, spaced as indicated under 2.2.1. On the other hand, if there is distortion present, its magnitude may be determined by making suitable spatial measurements of the optical image produced by the monitor, or by comparing the time pattern with a corresponding space pattern.

If blanking is present along with the timing signal, the aspect ratio may be determined.

2.2 *Practical Measuring Devices*

The apparatus and methods to be described are directly applicable to measurement of picture monitors using the present U.S. standards for monochrome television. The same general approach is obviously appropriate for other standards, however, when suitable quantitative changes are made.

2.2.1 Linearity Chart: The Monitor Linearity Chart, Fig. 1, illustrates a typical linearity chart used for checking aspect ratio and geometric distortion of a

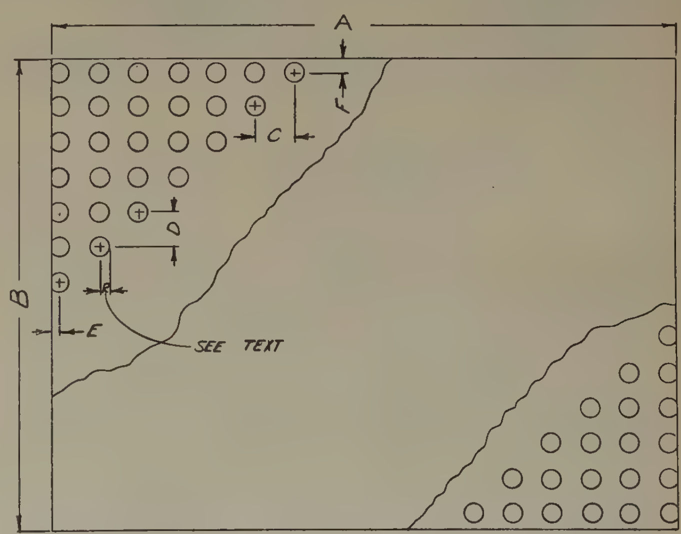

Fig. 1—Monitor linearity chart.

monitor. The size, shape, and physical form of the chart actually used will depend upon the physical characteristics of the monitor. The relative diameter of the circles depends upon the limits of distortion which are allowable in any given instance.

The relative dimensions, expressed in terms of frame height (for use with the procedures to be described below) are as follows:

Width—(Dimension "A")	4/3
Height—(Dimension "B")	1
Separation of circle centers horizontally—(Dimension "C")	0.081
Separation of circle centers vertically—(Dimension "D")	0.072
Separation of circle center from right or left edge—(Dimension "E")	0.018
Separation or circle center from top or bottom—(Dimension "F")	0.032

The radius of the circles should be made equal to the position error which is of interest expressed as a fraction of frame height. If desired, more than one circle may be described about each center point.

A reproduction of the RETMA chart shown in Fig. 3[1] (opposite page) may also be used.

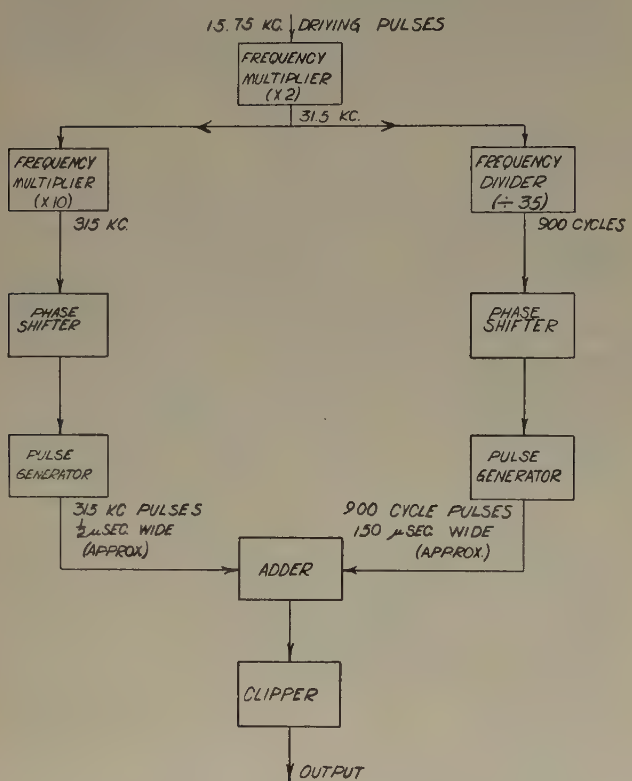

Fig. 2—Block diagram time-pattern generator.

2.2.2 Time-Pattern Generator: Fig. 2 shows a block diagram of a time-pattern generator. In order to produce an electrical grating pattern suitable for use with the chart described above, pulses having repetition rates of 315 kilocycles and 900 cycles must be used, representing values which are 20 times the horizontal sweep rate and 15 times the vertical sweep rate, respectively. This would give an array of 20 by 15 lines or dot-rows, were it not for the blanking intervals. The generator may be so designed as to give either lines or dots as a time pattern, depending upon the preference of the user. The nature of the pattern will be determined by whether the negative or positive portion of the waveform coming from the adder (Fig. 2) is retained by the clipper.

[1] See Part II, 4.2.1.

The 315-kilocycle pulses should preferably be less than 0.5 microsecond in duration. The display of the 900-cycle pulses should correspond to that of the 315-kilocycle pulses, the requirement on duration being 150 microseconds (about 2.5 lines) or less. In this manner, a suitable time pattern will be obtained.

2.3 Requirements and Characteristics of Measuring Equipment

2.3.1 Monitor Linearity Chart: The chart shown in Fig. 1 should be reproduced upon some convenient transparent material, the viewed area of which has the size and shape of the visible raster which is to be produced by the monitor. Suitable provisions must be made to avoid parallax errors and to keep the material plane. This may be done either by mounting the chart very close to the optical image, or by marking both sides of the transparent chart material in such a way that the circles on one side appear exactly opposite those on the other. The other physical characteristics should be made such as to allow the chart to be placed over the monitor viewing screen and held in place.

An alternative method of optical superposition is to use an opaque linearity chart at right angles to the monitor tube face in such a position that both the chart and the tube face can be seen simultaneously through a half-silvered mirror. This mirror is located directly in front of the tube and at an angle of 45 degrees with the face. Thus, the dot pattern on the monitor tube face is seen by transmission through the mirror and the chart by reflection.

2.3.2 Time-Pattern Generator: The position of the horizontal and vertical lines or dot-rows of the time-pattern generator should be made adjustable; this may be done by means of suitable phase shifters, as suggested by Fig. 2. Any other method will be satisfactory, if it meets the requirements that both sets of lines or dot-rows be movable for a distance at least equal to their separation.

This time-pattern signal should preferably have the blanking signal added in such a way that the output voltage corresponds to "black" during the blanking periods. The relative polarities of the time pattern and the blanking signal will, of course, determine whether the pattern background will appear in black or white on the monitor. The exact method of accomplishing these ends will depend to a great extent upon the design of the particular equipment which is being used.

3. Procedure

The methods outlined in this procedure may be used not only for measurement, but also for adjustment of the deflection system of the picture monitor.

3.1 Conditions

The picture monitor focus, contrast, and brightness controls should be adjusted so as to give a sharp, bright image when the proper video signals are applied.

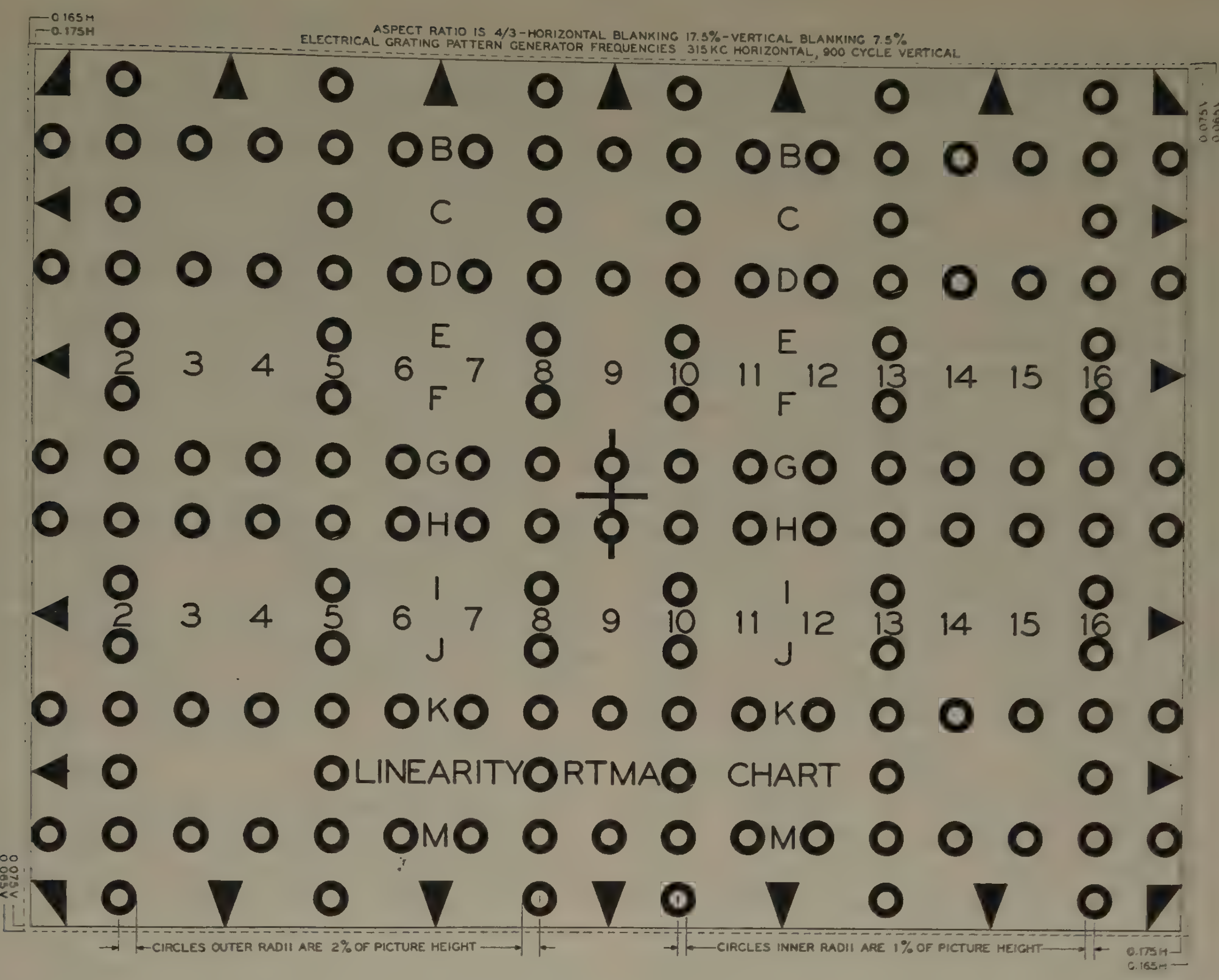

Fig. 3—Linearity RETMA chart.

3.1.1 Blanking: For measurements using the chart shown in Fig. 1, the vertical blanking should be set to 7.5 per cent, and the horizontal to 17.5 per cent, to obtain proper conditions for measuring and setting the aspect ratio. (This assumes measurements are to be made on a system using the present U. S. standards for monochrome television.)

3.2 Measurement Technique

3.2.1 General: The time-pattern signal should be connected to the monitor input. Level of the crosshatch signal with respect to blanking should be set to give a convenient value of setup. Deflection amplitudes should be set to the proper nominal values, and centering adjusted to give the proper raster position. The chart is then placed over monitor viewing screen.

3.2.2 Monitor Aspect Ratio: This procedure may be more properly described as an adjustment rather than a measurement. The proper aspect ratio of 4/3 will be obtained when the raster size is so adjusted as to fill exactly the area defined by the boundary on the chart in Fig. 1, provided vertical blanking has been set to 7.5 per cent, and the horizontal to 17.5 per cent.

3.2.3 Monitor Geometric Distortion: In order to determine geometric distortion, the time-pattern lines or dot-rows should be moved horizontally and vertically by means of the phasing controls until the dots or the intersections of the lines fall as close as possible to the centers of their corresponding circles on the chart. If the chart has been marked both front and rear, parallax errors may be avoided by viewing it so that the rear circles are exactly covered by the front circles. (This may or may not be possible for the entire area at once. The geometric position error at any point is then given by the magnitude of the distance by which an intersection or dot lies from its *corresponding* circle center If the circles have been made to have a radius equal to a certain percentage of frame height, then, if an intersection or dot falls within its corresponding circles, the geometric distortion must be less than that percentage. Care should be taken to assure that they are observed with respect to their *corresponding* circles.

If, as a result of this measurement, an attempt is made to adjust the deflection circuits for less geometric distortion, care should be taken to maintain the aspect ratio.

3.3 Presentation of Data

With the above-described measurement, it is possible to state over how much of the area of the picture and in what parts of the raster the position error exceeds a certain value. The facts may be presented in any convenient manner.

Part II

4. Measurement of Camera

4.1 Apparatus and Circuits

4.1.1 Basic Methods: The most convenient method of measuring geometric distortion is to measure the position of the aperture at regular intervals of time. These intervals can best be marked by means of suitable periodic electrical signals. In particular, two sets of intervals are used, each representing some fraction of the horizontal and vertical deflection periods, respectively. When this electrically generated signal is used in place of the picture signal, a regular pattern of lines or dots is produced on the picture monitor, which is called the time pattern. For checking geometric distortion in the camera, a space pattern consisting of a special test chart is used. This chart consists of an array of regularly-spaced markers. The number of markers in either direction is made to coincide with the number of electrical markers in the time pattern. The picture signal from the camera having this space pattern in its field of view is then added to the time pattern. This signal is then applied to any convenient picture monitor, permitting the comparison of the two patterns. It should be noted that this method does not depend upon the use of a picture monitor which is free from geometric distortion.

If the proper blanking is present, the space-pattern chart may also be used to adjust the aspect ratio of the camera.

4.2 Practical Measuring Devices

The apparatus and methods to be described are directly applicable for measuring television systems using the present U.S. standards for monochrome television. The same general methods are obviously appropriate for other standards, however, when suitable quantitative changes are made.

4.2.1 Camera Linearity Chart (Space Pattern): A chart has been developed by the RETMA for the purpose of checking the geometric distortion and adjusting the aspect ratio of the camera. A reproduction of this chart is shown in Fig. 3.[2] Suitable slides are easily made from such a chart. The circles have an inner radius equal to one per cent of the frame height and an outer radius equal to two per cent of the frame height. Provisions are made for defining the edges and corners of the chart accurately; an array of letters and numbers is used to help identify rows and columns of circles.

4.2.2 Time-Pattern Generator: Fig. 2 shows a block diagram of a time-pattern generator. In order to produce and electrical grating pattern suitable for use with the chart described above, pulses having repetition rates of 315 kilocycles and 900 cycles must be used, representing values which are 20 times the horizontal sweep rate and 15 times the vertical sweep rate, respectively. This would give an array of 20 by 15 lines or dot-rows, were it not for the blanking intervals. The generator may be so designed as to give either lines or dots as a time pattern, depending upon the preference of the user. The nature of the pattern will be determined by whether the negative or positive portion of the waveform coming from the adder (Fig. 2) is retained by the clipper.

The 315-kilocycle pulses should preferably be less than 0.5 microsecond in duration. The display of the 900-cycle pulses should correspond to that of the 315-kilocycle pulses, the requirement on duration being 150 microseconds (about 2.5 lines) or less. In this manner, a suitable time pattern will be obtained.

4.3 Requirements and Characteristics of Measuring Equipment

4.3.1 Camera Linearity Chart: The chart should be reproduced in that size and form which is appropriate for the camera whose geometric distortion is to be measured. The standard 18-inch by 24-inch chart will be found to be suitable for direct pickup cameras. For film cameras, a reproduction of this chart is necessary, usually in the form of a transparency which will fit the projector associated with the camera to be measured. The size and position of the space pattern on the transparency should be such as to correspond to the area which will be scanned (during the unblanked period) when the camera is in normal use.

4.3.2 Time-Pattern Generator: The position of the horizontal and vertical lines or dot-rows of the time pattern generator should be made adjustable; this may be done by means of suitable phase shifters, as suggested by Fig. 2. Any other method will be satisfactory, if it meets the requirements that both sets of lines or dot-rows be movable for a distance at least equal to their separation.

This time-pattern signal preferably should have the blanking signal added in such a way that its output voltage corresponds to "black" during the blanking periods. The relative polarities of time pattern and blanking signal will determine whether the pattern background will appear in black or white on the monitor. Also suitable provisions should be made for adding this time-pattern signal to picture-signal output of the camera under measurement. The exact method of accomplishing these ends will depend to a great extent upon the design of the particular equipment which is being used.

[2] The original is 18 inches high and 24 inches wide and available from the RETMA Data Bureau.

5. Procedure

5.1 Conditions

The camera and picture monitor should be so adjusted as to give a good picture, at least from the point of view of the picture signal itself (i.e., without respect to geometric distortion). The methods outlined in this procedure may be used not only for measurement, but for adjustment of deflection system of the camera.

5.1.1 Blanking: For measurements using the chart shown in Fig. 3, the vertical blanking should be set 7.5 per cent, and the horizontal to 17.5 per cent, to obtain proper conditions for measuring or setting the aspect ratio. (This assumes measurements are to be made on a system using the present U.S. standards for monochrome television.)

5.1.2 Picture Monitor Deflection Adjustment: The monitor should preferably be adjusted as well as possible for correct aspect ratio and minimum geometric distortion, in accordance with the methods outlined in Part I. It is highly desirable to have the raster size on the monitor so adjusted, for the purposes of camera measurement, that it is visible in its entirety; this may mean setting it smaller than normal.

It should be noted that the accuracy with which the monitor is adjusted does not limit the accuracy with which the camera may be measured, with the method to be described. It is, nevertheless, convenient to have the monitor as well-adjusted as possible.

5.2 Measurement Technique

5.2.1 General: With the chart of Fig. 3 placed before the camera, the size and position of the camera raster should be set to their proper nominal values. The optical image size is then adjusted, in the case of the direct pickup camera, by changing the distance between the camera and the chart, making certain to maintain optical focus. Similarly, the optical position may be adjusted by suitable movement of chart or camera. When in proper final position, the axis of the camera lens must be perpendicular to the chart and pass through its center.

In film cameras, adjustment of the optical image is not usually possible; it should be unnecessary, however, if the slide has been properly prepared in accordance with paragraph 4.3.1.

5.2.2 Camera Aspect Ratio: The following procedure may be more properly described as an adjustment rather than a measurement. It may be convenient in this procedure to have the time-pattern signal removed. With the camera raster having the correct nominal size and position, its exact size and position are adjusted until all four edges of the chart, as designated by the points of the arrowheads, coincide as nearly as possible with edges of the raster, as viewed on the picture monitor.

The dotted lines outside the solid frame boundary of the chart are intended for scaling purposes only and are to be ignored in the measurement.

If this camera is adjusted as described above, its picture signal will meet the specifications of a 4-to-3 aspect ratio provided the vertical blanking has been set to 7.5 per cent and the horizontal to 17.5 per cent.

5.2.3 Camera Geometric Distortion: For this measurement, the time-pattern and picture signals are viewed simultaneously on the picture monitor. The conditions described in 5.2.2 must be maintained at all times. The lines or dots are then moved, by means of the phasing controls, horizontally and vertically, until the dots or line intersections fall within as many as possible of the circles of the chart. If a dot or line intersection falls within its *corresponding* circle, the position error at that point is one per cent or less of the picture height; if it falls on the dark portion of its corresponding circle, the position error at that point is between one and two per cent of the picture height; if it falls outside of its corresponding circle, the position error at that point is over two per cent. In any case, the amount of position error may be estimated by comparing its displacement with the size of the circles.

It should be noted that the dot or line intersection must be measured with respect to its corresponding circle; e.g., the time marker on column 2, row 3, should be measured with respect to that circle marked number 2 and letter *C*.

It may be of value to note the *direction* of the geometric distortion, especially on adjacent rows or columns of circles. Thus, for example, the error on the second column of circles might be to the left, while that of the third column might be to the right. In both cases, the value of the position error might be within the desired tolerance; nevertheless, the sudden shift of the error from one extreme to the other may indicate the presence of rapid velocity changes in the scanning, which are more evident with a moving image than with a static test such as herein described.

If as a result of this measurement an attempt is made to adjust the camera deflection circuits for less geometric distortion, care should be taken to maintain the aspect ratio.

5.3 Presentation of Data

With the above-described measurement, it is possible to state over how much of the area of the picture and in what parts of the raster the position error exceeds a tolerance of either one or two per cent. These facts may be presented in any convenient manner.

It should be noted that the numbers and letters, which identify the rows and columns of the circles on the chart, may be used as an aid in specifying areas or locations of interest in presenting the data.

Rocket Instrumentation for Reliable Upper-Atmosphere Temperature Determination*

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Summary—**This paper describes briefly rocket-borne electronic equipment which has been developed and utilized in the determination of ambient atmospheric pressure and temperature. In the design of the equipment, emphasis has been placed on reliability of operation, and on the ability to produce data of significant accuracy. Typical resulting curves of ambient pressure and temperature are presented.**

Introduction

The DESIRE to obtain measurements of the ambient temperature, density, and pressure of the upper atmosphere with significant accuracy has led to the development of equipment which produces rather precise data dependably.

During the exploratory phase of upper-atmosphere temperature measurement utilizing rockets following World War II, the temperature was determined through the use of the barometric equation, a computational procedure which involves, fundamentally, differentiation of atmospheric pressure as a function of altitude to obtain temperature as a function of altitude. This method tends to exaggerate the errors of the original pressure measurement and leads generally to temperatures having accuracies of 1 part in 10. Due to this and other shortcomings, this method has been largely abandoned by this research group.

Newer methods,[1,2,3,4] have been developed which currently yield temperatures having accuracies of approximately 1 part in 50. One of these[1] utilizes the ratio of two pressures, one at the tip of a rocket nose-cone and the other some distance aft of the tip on an element of the cone, in a determination of a Mach number, which leads, when considered with missile velocity, to ambient temperature.

It is the purpose of this paper to describe briefly the equipment which has implemented this method. The instrumentation consists essentially of pressure-measuring instruments, but includes equally important accessory equipments whose primary function permits (a) computation of the space position of rocket as a function of time, (b) photographic recording of data in missile, (c) determination of angular orientation of missile, and (d) calibration of data-telemetering system.

The total system, including the basic pressure-sensitive devices, is fundamentally electronic in nature.

Considerations in Equipment Design

Two important considerations are necessary to the design and development of unmanned rocket-borne electronic instrumentation: first, that the equipment be capable of performing the desired function under normal anticipated flight conditions of high velocity, acceleration, vibration, missile surface temperature and low ambient pressure; and second, less tangible, that the equipment perform satisfactorily under numerous possible anomalous conditions of flight including extreme shock or vibration, electrical noise, excessive yaw or roll rate, and change of data-recording system calibration.

Reliability of operation is paramount, for once the rocket leaves the ground, a "point of no return" is reached. Subsequent readjustment or correction of possible equipment malfunction is not possible, and any failure represents an irremediable and costly loss.

Simplicity of design and construction contributes significantly to reliability and thus constitutes a controlling consideration in equipment planning.

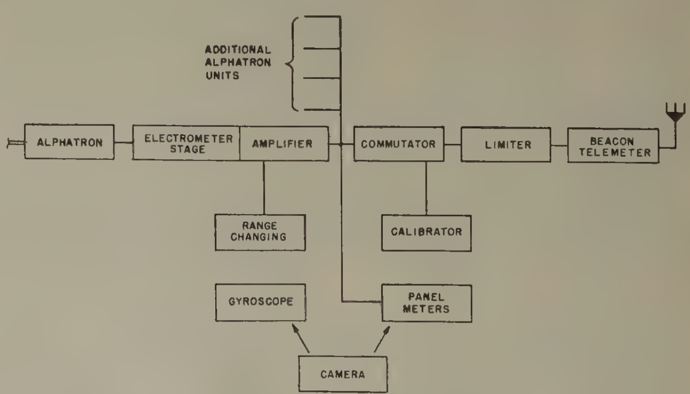

Fig. 1—Simplified block diagram of typical Aerobee rocket instrumentation

System Outline

Fig. 1 is a simplified block diagram of a pressure-measurement instrumentation system for an Aerobee rocket. Only one alphatron pressure gage with its associated amplifier and range-changing circuit is shown,

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[1] H. S. Sicinski, N. W. Spencer, and W. G. Dow, *Jour. Appl. Phys.*, vol. 25; February, 1954.

[2] M. J. Ference, "Temperature in earth's upper atmosphere," *Trans. Amer. Geophys. Union*, vol. 33, p. 317; 1952.

[3] L. M. Jones and F. L. W. Bartman, Eng. Res. Inst., Univ. of Michigan, Ann. Arbor, Mich; Report "Falling Sphere Method for Upper Air Density and Temperature"; August, 1953.

[4] R. J. Havens, R. T. Koll, and H. E. Lagow, *Jour. Geophys. Res.*, vol. 57; March, 1952.

although a typical instrumentation would contain at least five similar independent units. One of these five measures the pressure at the cone tip, the other four the pressures at various points on the cone surface. The remaining equipment shown in the block diagram is in general not duplicated in the system.

The general measurement plan is as follows: the pressure signal appears in the alphatron gage through appropriate vacuum plumbing from a port on the rocket nose-cone. The gage, amplifier, and range-changing circuit act together to translate the pressure to a voltage which is in turn presented to the telemeter through the commutator and limiter. The telemeter, which also serves as a beacon, transmits the desired information to three ground stations where decoding and recording is accomplished.

The beacon-telemeter transmits in response to ground-based interrogation, the replies being received at the three ground stations, which are appropriately located geographically to permit missile-position determination through the use of a triangulation computational procedure.

Alphatron System

An alphatron[5] is an ionization gage wherein a small quantity of radium provides the ionizing energy. Alpha particles emanating from the radium source have sufficient energy to cause many ionizations on collision with gas particles in the gage chamber. Suitably polarized electrodes maintain an electric field to effect collection of the ionization products. The resulting current in the external circuit is proportional to the number of ionized particles produced per second and hence to the density of gas in the chamber.

It has been determined by measurement that the temperature of the walls of the gage chamber, and hence also, of the gas, does not change significantly during the short period of the rocket flight; therefore for the density interval considered, the ratio of gas pressure to density is constant, and the alphatron gage may be calibrated in terms of gas pressure.

Fig. 2 illustrates a typical alphatron functionally, showing the relative locations of the electrodes and the radium source. Fig. 3 is a typical curve of output current versus chamber pressure, for constant temperature.

It may be noted that the response curve is not linear throughout the range shown. The departure from linearity at the higher pressures is due to recombination of some of the ions and electrons before collection can occur. The departure from linearity at the lower pressures is due to a so-called "dark current," whose cause has not been firmly established. Alpha particles (helium nuclei) bombard the collector with sufficient energy to cause secondary emission; also, the alpha particles may terminate at the collector and be neutralized. These two influences, not essentially density-dependent, presumably contribute to the establishment of the dark current. The low-pressure portion of the gage characteristic is asymptotic to the dark current.

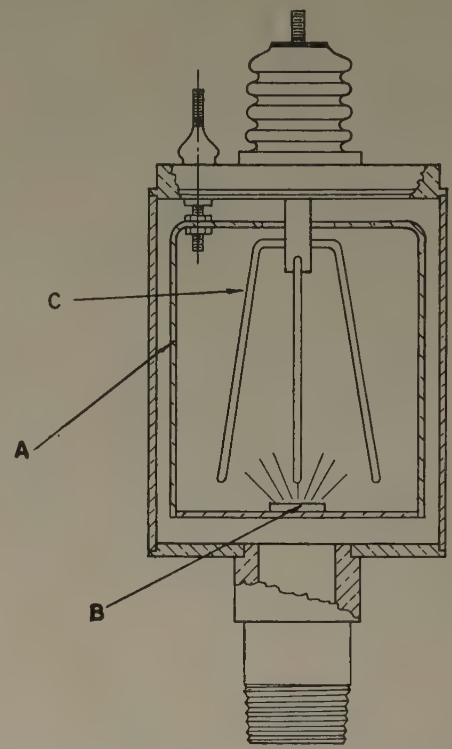

Fig. 2—Diagram of alphatron pressure gage: (a) electrode, (b) radium source, (c) collector electrode.

In order to utilize the alphatron to advantage, associated circuits must accommodate the minute Fig. 3 alphatron current, and, depending on the requisite pressure resolution, subdivide the total useful range into appropriate subranges. In addition, the alphatron cur-

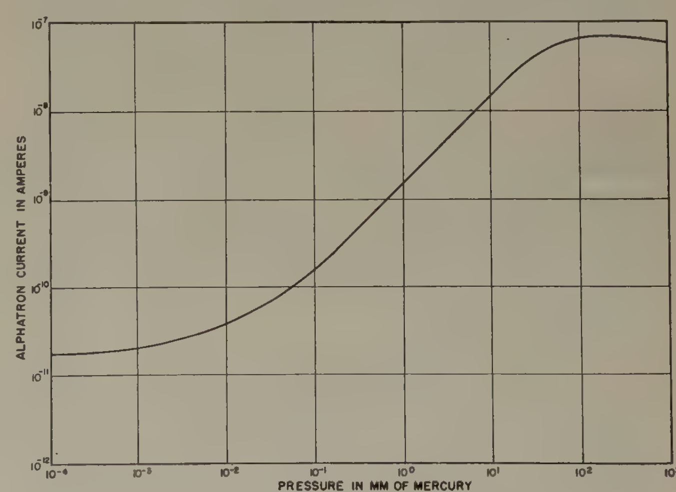

Fig. 3—Typical alphatron-pressure-gage characteristic for constant temperatures.

rent over each subrange must be translated into voltages in accordance with the 0–5-volt dc range the telemeter will accept. The necessary pressure resolution sets the requirement that the total range be di-

[5] J. R. Downing and G. Mellen, "A sensitive vacuum gage with linear response," *Rev. Sci. Inst.*, vol. 17, pp. 218–223; June, 1946.

vided into eight subranges. For example, subrange 6 accommodates pressures in the range 0.06 to 0.005 mb, subrange 5 the range 0.3 to 0.06 mb, subrange 4 the range 1.23 to 0.3 mb, etc., each subrange corresponding to the output voltage range of 0–5 volts dc.

The alphatron acts like a current generator. Thus, in order to produce the correct voltage range for a given pressure range, one may provide appropriate values of load resistance, different for each subrange. It follows then that the various subranges may be distinguished physically in the system as different resistors of appropriate values. Because of the very low current values, large values of resistance are required, in some cases as high as 10^{11} ohms.

The voltage developed across the large value of resistance is applied to a dc amplifier which employs unity negative feedback. Accordingly, a system is obtained which has an extremely low source impedance and consequently the magnitude of the output or signal voltage is relatively unaffected by the loading that the data-recording systems impose. Due to the large value of resistance at the input of the amplifier, an electrometer input stage is employed to help minimize shunt leakage.

Because of the unity feedback, a change in input voltage (current change in the alphatron load resistance) produces a very nearly equivalent change in output voltage, independent of normal and anticipated variations of (a) vacuum-tube characteristic and (b) all resistances except the alphatron load resistance. Any change in the instantaneous dc level (drift) due to change of (a) cathode temperature and (b) battery supply voltage etc. is not affected by the use of feedback. Care in design and the use of conventional techniques for compensation of cathode temperature variation limit the drift to a negligible amount for the interval of the rocket flight. Thus calibration of the system is a function essentially of the alphatron and its load resistance.

Although the voltage gain of the system is unity, the current gain is of course very high, being numerically the ratio of the alphatron load resistance to the cathode resistor (5,000 ohms) of the amplifiers cathode-follower output stage. This relationship holds as long as the gain without feedback (in this case about 450) is very high compared with unity.

Range Changing

The telemeter will not relay voltages correctly outside the range 0 to +5 volts dc, although the amplifier system is linear to higher positive and through zero to negative voltages. The values of alphatron load resistances have been proportioned so that if the pressure changes to decrease the output signal to 0.4 volt, the next resistor, if inserted, will produce a system output voltage of 4.6 volts. Or conversely, if the output voltage goes to 4.6 volts, substitution of the next resistor in the other direction will cause a system output of 0.4 volt.

Thus, ideally, if the proper value of alphatron load resistance is inserted at the correct time by a suitable switching device, the system output signal will go from 0.4 volt to 4.6 volts, then return to 0.4 volt, increase again to 4.6 volts, and so forth as the pressure is varied throughout the total range. The reverse procedure holds equally well.

The range-changing circuit provides the requisite automatic switching function. It is a form of dc servomechanism which utilizes the amplifier output signal as an error signal. As the error signal reaches 0.4 (or 4.6) volts, the circuit acts through a rotary solenoid switch to substitute the next value of alphatron load resistance in the amplifier input circuit.

Operationally, the complete pressure-measuring system delivers to the telemetering system a voltage corresponding to a pressure measurement of accuracy approaching 1 part in 100 over the total desired range of atmospheric pressure to 10^{-3} mm Hg, nearly six decades of pressure changes. Correspondingly, pressure values recorded at ground stations have an accuracy of about 1 part in 65.

Commutator

A commutator (mechanical multiplexer) is provided to allow more efficient utilization of the two available telemeter channels. Thus a variety of information is conveyed by a single telemetering channel, including generally outputs of each alphatron unit incorporated in the instrumentation, a voltage indicating subrange in use by each unit, calibration voltages, and various possible signals relating to equipment functioning.

A typical commutator samples twenty different points, its output comprising a series of voltage pulses of various magnitudes. The scanning rate is generally set to produce approximately five complete sets of data per second.

Limiters

All voltages presented to the telemetering system for recording may be controlled by the limiters. In principle, the limiters, which employ biased thermionic or semiconductor diodes, function to prevent the application of any voltage to the telemeter that lies outside the range of 0 to +5 volts dc. This provision protects the telemeter from disturbing over-voltage or negative-voltage signals.

In view of the many different signals (20–40) presented to the telemeter, any equipment failure which could cause one of the signals to fall outside the normal anticipated range would disturb the telemeter circuits, and could conceivably cause erroneous recording of other valid information. Thus limiter protection is a necessary feature for reliable data transmission.

Gyroscope and Camera

A computation of temperature from pressure measurements requires knowledge of the rocket's attitude in relation to the air stream velocity vector. A gyroscope is included in the instrumentation to provide indications of two quantities which when considered with the missile's trajectory permit a determination of (a) the angle between missile's longitudinal axis and stream velocity vector and (b) roll position of the missile.

The gyro used requires that the data be recorded by camera; therefore, a 16-mm motion-picture camera, suitably modified for flight use, is carried in the rocket.

It was noted that the desired pressure information is presented to the telemeter for transmission to the ground. It is presented also on panel meters in the missile for recording, by the above camera, simultaneously with the gyroscope data.

The gyroscope is powered from batteries through the use of a vibrator-type inverter which provides 110-v, 400-cycle, single-phase power.

Beacon-Telemeter

The beacon telemetering unit, developed at the Air Force Cambridge Research Center, performs three functions necessary for the success of the rocket flight:

1. Transmission of experimental data to ground stations for recording.
2. Transmission of a signal to the ground which permits determination of the space position of the rocket.
3. Reception of signals from the ground, which permits control of rocket or experimental functions.

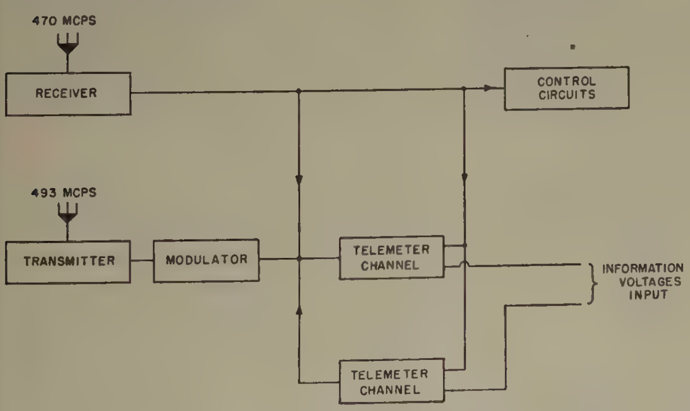

Fig. 4—Simplified block diagram of beacon-telemeter system.

The block diagram, Fig. 4, illustrates the relationship between the various portions of this equipment. Pulses from a ground-based interrogator at a frequency of 470 mc are received by the rocket-borne beacon, where they become triggers for the modulator-transmitter, which in turn replies at a frequency of 493.5 mc. The replies, known as range pulses, are received by three widely separated (20–40 miles) ground stations where the oscilloscope-displayed time delay between the interrogation and the range pulses is photographed for later use in rocket trajectory determination.

The interrogation pulses are utilized also as triggers for the telemetering channel portions of the beacon equipment which, in turn, produce delayed pulses to trigger the modulator. The delay of these "telemeter" pulses is under the control of the information voltages being applied in this case by the pressure-measuring instrumentation. For example, for 0 volts applied, the channel 1 telemetering pulse is delayed 50 microseconds after the range pulse. For 5 volts applied this delay becomes 200 microseconds. Channel 2 has corresponding but greater delays.

Thus a single interrogation pulse produces a range pulse whose delay after the time of the interrogation indicates the distance of the rocket from the ground station, plus a pulse for each telemetering channel whose delay with reference to the range pulse indicates the information voltage being applied. The three pulses when received at a ground station are applied as intensity modulation to an oscilloscope whose time base is triggered by the range pulse of the set. Photography of the oscilloscope provides a record of the flight voltages.

The control functions mentioned above are performed by two-relay-operating circuits, one of which is sensitive to the repetition frequency of the interrogation, the other to the presence or absence of interrogation. Therefore, circuits and the functions they perform are under control of the ground station operator.

Conclusions

The foregoing sections serve to indicate, in keeping with the scope of this paper, the major component equipments employed in a typical instrumented Aerobee rocket. Instrumentation for obtaining measurements other than pressure will differ in detail but will, in most cases, be subject to similar considerations and limiting requirements.

There are, of course, many accessory devices not mentioned in this paper, whose functions complement and facilitate operation of the equipment discussed. Included among these items are varied remote-control facilities for amplifier adjustment prior to flight, thermometric devices for equipmental or missile spot-temperature determination, acceleration indicators, camera-film time-base generators, film-run indicators, gyro-current indicators, various timing devices, data-recording system calibrators, and, of course, primary power-supply battery systems.

Instrumentation of the kind described has been used on a number of Aerobee rocket flights. In all these flights the equipment described has functioned as intended. Evaluated in terms of reliability and capacity to yield valid results, the equipment is believed to be a significant contribution to electronic instrumentation.

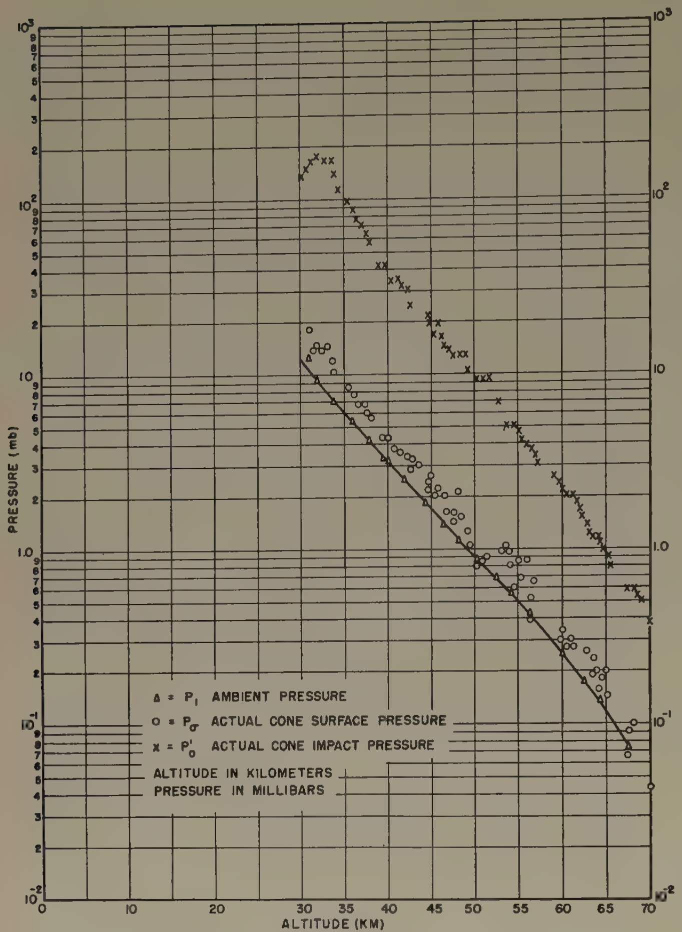

Fig. 5—Air pressures measured by instrumentation in an Aerobee rocket for September 13, 1951.

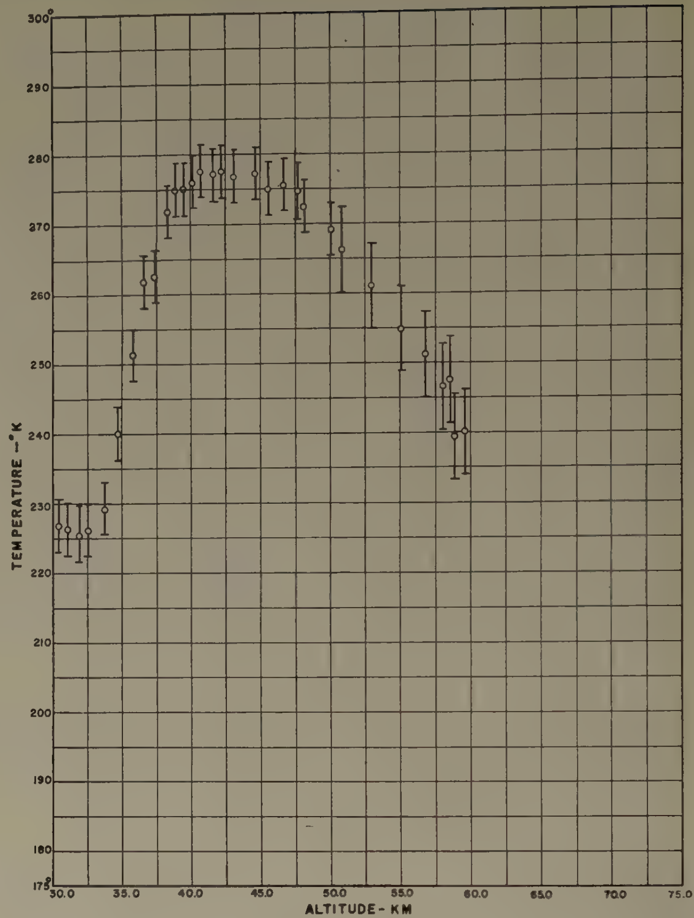

Fig. 6—Ambient air temperature computed from air pressures measured from an Aerobee rocket for September 13, 1951.

Similarly, resulting upper-atmosphere pressure, temperature, and density information comprise a significant portion of presently existing data obtained by direct measurement, which high-altitude rockets permit.[6]

[6] "Rocket Panel," *Phys. Rev.*, vol. 88, pp. 1027–1032; Dec., 1952.

Typical experimental results which were obtained during a particular field test, conducted September 13, 1951, are illustrated in Figs. 5 and 6, which present the ambient atmospheric pressure and temperature for that date.

IRE Standards on Audio Techniques: Definitions of Terms, 1954*

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I. Introduction

This standard establishes concise meanings for such general audio transmission terms as gain, loss, amplification, and attenuation, to permit standardization of methods of measurement. After review of the conflicting past usage in the communication art, the committee sponsored this set of meanings that establishes basic and logical relationships for both analytical and laboratory use among the general terms and allows for future expansion in application.

It should be noted that amplification and attenuation are defined as the numerical ratio of any two magnitudes of any quantity, for instance, current or voltage; while gain and loss are so defined as to restrict them to power ratios only. The expression of these concepts in ratio form makes it possible to define for a linear system or component a characteristic independent of the value of a quantity such as power, current, or voltage.

* Reprints of this Standard, 54IRE3.S1, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy. A 20-per cent discount will be allowed for 100 or more copies mailed to one address.

II. Definitions

Active Transducer

A transducer whose output waves are dependent upon sources of power, apart from that supplied by any of the actuating waves, which power is controlled by one or more of these waves.

Amplification

General transmission term used to denote an increase of signal magnitude.

Attenuation

General transmission term used to denote a decrease of signal magnitude.

Available Power (of a Linear Source of Electric Energy)

The quotient of the mean square of the open-circuit terminal voltage of the source divided by four times the resistive component of the impedance of the source.

Bridging

The shunting of one electrical circuit by another.

Bridging Gain

The ratio of the power a transducer delivers to a specified load impedance under specified operating conditions to the power dissipated in the reference impedance across which the input of the transducer is bridged.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This gain is usually expressed in decibels.

Bridging Loss

The ratio of the power dissipated in the reference impedance across which the input of a transducer is bridged, to the power the transducer delivers to a specified load impedance under specified operating conditions.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This loss is usually expressed in decibels.

Note 3: In telephone practice this term is synonymous with the insertion loss resulting from bridging an impedance across a circuit.

Current Amplification

The ratio of the magnitude of the current in a specified load impedance connected to a transducer to the magnitude of the current in the input circuit of the transducer.

Note 1: If the input and/or output current consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: By custom this amplification is often expressed in decibels by multiplying its common logarithm by 20.

Current Attenuation

The ratio of the magnitude of the current in the input circuit of a transducer to the magnitude of the current in a specified load impedance connected to the transducer.

Note 1: If the input and/or output current consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: By custom this attenuation is often expressed in decibels by multiplying its common logarithm by 20.

dbm

A symbol for power level in decibels with reference to a power of one milliwatt (0.001 watt).

Gain (Transmission Gain)

General term used to denote an increase in signal power in transmission from one point to another. Gain is usually expressed in decibels and is widely used to denote Transducer Gain.

Ideal Transducer (for Connecting a Specified Source to a Specified Load)

A hypothetical passive transducer which transfers the maximum possible power from the source to the load.

Note 1: In linear transducers having only one input and one output, and for which the impedance concept applies, this is equivalent to a transducer which (a) dissipates no energy and (b) when connected to the specified source and load presents to each its conjugate impedance.

Ideal Transformer

A hypothetical transformer which neither stores nor dissipates energy. Its self inductances have a finite ratio and unity coefficient of coupling. Its self and mutual impedances are pure inductances of infinitely great value.

Input Impedance

The impedance presented by the transducer to a source.

Insertion Gain

Resulting from the insertion of a transducer in a transmission system, the ratio of the power delivered to that part of the system following the transducer to the power delivered to that same part before insertion of the transducer.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This gain is usually expressed in decibels.

Note 3: The "insertion of a transducer" includes bridging of an impedance across the transmission system.

Insertion Loss

Resulting from the insertion of a transducer in a transmission system, the ratio of the power delivered to that part of the system following the transducer, before insertion of the transducer, to the power delivered to that same part of the system after insertion of the transducer.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This loss is usually expressed in decibels.

Note 3: The "insertion of a transducer" includes bridging of an impedance across the transmission system.

Level

The difference of a quantity from an arbitrarily specified reference quantity.

Note 1: In Audio Techniques, the quantities of interest are often expressed in decibels, thus their difference is conveniently expressed as a ratio. Hence, level is widely regarded as the ratio of the magnitude of a quantity to an arbitrary reference magnitude.

Load

(1) The device which receives signal power from a transducer.

(2) The signal power delivered by a transducer (deprecated).

Load Impedance

The impedance presented by the load to a transducer.

Loss (*Transmission Loss*)

General term used to denote a decrease in signal power in transmission from one point to another. Loss is usually expressed in decibels.

Output Impedance

The impedance presented by the transducer to a load.

Passive Transducer

A transducer whose output waves are independent of any sources of power which are controlled by the actuating waves.

Power Amplification

See *Power Gain.*

Power Attenuation

See *Power Loss.*

Power Gain

The ratio of the power that a transducer delivers to a specified load, under specified operating conditions, to the power absorbed by its input circuit.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This gain is usually expressed in decibels.

Power Level

At any point in a transmission system, the difference of the measure of the steady state power at that point from the measure of an arbitrarily specified amount of power chosen as a reference.

Note: In Audio Techniques, the measures are often expressed in decibels, thus their difference is conveniently expressed as a ratio. Hence, power level is widely regarded as the ratio of the steady state power at some point in a system to an arbitrary amount of power chosen as a reference.

Power Loss

The ratio of the power absorbed by the input circuit of a transducer to the power delivered to a specified load under specified operating conditions.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This loss is usually expressed in decibels.

Program Level

The measure of the program signal in an audio system expressed in vu.

Program Signal

In Audio Systems and components, the complex electric wave corresponding to speech, music, and associated sounds, destined for audible reproduction.

Reference Volume

The volume which gives a reading of 0 vu on a standard volume indicator.

Signal

(1) A visual, audible, or other indication used to convey information.

(2) The intelligence, message, or effect to be conveyed over a communication system.

(3) A signal wave.

Signal Level

At any point in a transmission system, the difference of the measure of the signal at that point from the measure of an arbitrarily specified signal chosen as a reference.

Note 1: In Audio Techniques, the measures of the signal are often expressed in decibels, thus their difference is conveniently expressed as a ratio.

Source

The device which supplies signal power to a transducer.

Source Impedance

The impedance presented by the source to a transducer.

Standard Volume Indicator

A device for the indication of volume having the characteristics prescribed in ASA-C16.5.

Transducer

A device capable of being actuated by waves from one or more transmission systems or media and of supplying related waves to one or more other transmission systems or media.

Transducer Gain

The ratio of the power that the transducer delivers to a specified load under specified operating conditions to the available power of a specified source.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This gain is usually expressed in decibels.

Transducer Loss

The ratio of the available power of a specified source to the power that the transducer delivers to a specified load under specified operating conditions.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This loss is usually expressed in decibels.

Transformer Loss

The ratio of the power that would be delivered to a specified load impedance if an ideal transformer were substituted for the actual transformer, to the power delivered to the specified load impedance by the transformer, under the condition that the impedance ratio of the ideal transformer is equal to that specified for the transformer.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This loss is usually expressed in decibels.

Transformer Loss (Deprecated)

The loss which would be eliminated by the insertion, at any point in a transmission system, of an ideal transformer having an impedance ratio equal to the absolute value of the ratio of the impedances facing the transformer.

Note 1: This loss is usually expressed in decibels.

Transition Loss (In Audio Systems and Components)

At any point in a transmission system, the ratio of the available power from that part of the system ahead of the point under consideration to the power delivered to that part of the system beyond the point under consideration.

Note 1: If the input and/or output power consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: This loss is usually expressed in decibels.

Voltage Amplification

The ratio of the magnitude of the voltage across a specified load impedance connected to a transducer to the magnitude of the voltage across the input of the transducer.

Note 1: If the input and/or output voltage consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: By custom this amplification is often expressed in decibels by multiplying its common logarithm by 20.

Voltage Attenuation

The ratio of the magnitude of the voltage across the the input of the transducer to the magnitude of the voltage delivered to a specified load impedance connected to the transducer.

Note 1: If the input and/or output voltage consist of more than one component, such as multifrequency signal or noise, then the particular components used and their weighting must be specified.

Note 2: By custom this attenuation is often expressed in decibels by multiplying its common logarithm by 20.

Volume

The magnitude of a complex audio frequency wave in an electric circuit as measured on a standard volume indicator. The volume is expressed in vu. In addition, the term volume is used loosely to signify either the intensity of a sound or the magnitude of an audio frequency wave.

vu (Pronounced "vee-you" and customarily written with lower case letters)

A quantitative expression for volume in an electric circuit.

Note 1: The volume in vu is numerically equal to the number of db which expresses the ratio of the magnitude of the waves to the magnitude of reference volume.

Note 2: The term vu should not be used to express results of measurements of complex waves made with devices having characteristics differing from those of the standard volume indicator.

The Application of Some Semiconductors as Logarithmic Elements*

N. M. SCHAEFFER†, AND G. W. WOOD‡, SENIOR MEMBER, IRE

Summary—Practical and theoretical considerations involved in the application of semiconductors as passive logarithmic elements are discussed. Simple voltage-divider circuits incorporating semiconductors exhibiting special characteristics are used to obtain output voltages proportional to the logarithms of input voltages, with emphasis on the dc case. Data are presented on the logarithmic response, and temperature dependence of a number of specific types of rectifiers. Reasonably good temperature compensation of such circuits can be achieved with negative temperature coefficient resistors. Circuits having a logarithmic response have been demonstrated for input signals from dc up to 100 kilocycles in frequency.

Introduction

Although research upon which this paper is based was performed under a contract with the Navy Department, Bureau of Ordnance, with a view toward a particular application, the information yielded is somewhat broader in scope. The initial objective was the development of a circuit which would give a dc output voltage proportional to the logarithm of the input voltage: over a range of input voltages from approximately 0.1 volt to 30 volts dc; operate between source and load impedances of approximately 1 megohm; consume no power except that derived from the signal source; operate with minimum error between temperatures of 30 degrees and 90 degrees F.; and have minimum insertion loss.

The applications for logarithmic circuits are many, particularly in electrical computing devices and in the display or recording of electrical quantities according to a decibel scale.

Several investigations of electrical circuits which give an output voltage proportional to the logarithm of the input voltage are reported in the literature.[1,2] In all cases the heart of such a circuit is essentially a nonlinear resistance. In some instances the nonohmic element is a vacuum tube or tubes, the use of which is not considered here because of the restriction that no power be consumed except from the signal. In other instances the nonohmic element is some type of semiconductor. The problem was soon resolved into a search for a suitable semiconductor, the current through which would vary exponentially with the applied voltage in such a way that a simple logarithmic voltage divider circuit could be used.

Kallman has shown that copper oxide rectifiers are suitable for circuits involving resistances of the order of a hundred ohms. Copper oxide is not as suitable for circuits containing the high resistances to which the present investigation is limited as are rectifiers of some other materials. This is shown in the data presented below.

Some of the more common nonlinear circuit elements are readily eliminated. The thermistor is sensitive to ambient temperature variations as well as to current-produced temperature changes which are not instantaneous. Thyrite is another nonlinear resistance material, the current through which is proportional to some power of the voltage drop across it. Both thermistors and thyrite are symmetrical in the sense that their characteristics are independent of the polarity of the applied potential. The materials giving the most promising results in the solution of the problem were unsymmetrical nonlinear resistances: germanium and silicon diodes, selenium and copper oxide rectifiers.

The theory of semi-conductors predicts a dc exponential current-voltage characteristic for the germanium diode such that in certain circuits the desired logarithmic relation between input and output voltages is obtained.[3] This circuit is shown in Fig. 1. In the discussion and data which follow, the series resistance R_S, and the load resistance R_L are each one megohm except where otherwise noted.

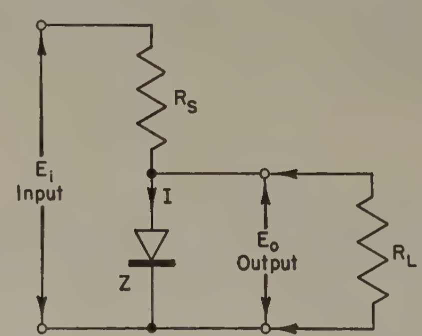

Fig. 1—Logarithmic voltage divider.

When both the source or series resistance and the load resistance are very large in comparison with the rectifier resistance, the output voltage is proportional to the logarithm of the input voltage. Similar relationships characterize selenium and copper-oxide disc rectifiers.

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[1] H. E. Kallmann, "Nonlinear circuit element applications," *Electronics*, vol. 19, p. 130; August, 1946.

[2] H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers," McGraw-Hill Book Co., Inc., New York, N. Y., p. 77; 1948.

[3] Torrey and Whitmer, *loc. cit.*

Experimentally, it has been found possible to fulfill the foregoing requirements over a wide range of input voltages, though for given source and load resistances the position of this range is dependent upon the rectifier resistance. The forward resistance of the rectifier is in turn dependent upon the material comprising the unit and its construction.

Preliminary Considerations

The parameters of a logarithmic voltage divider are subject to certain other restrictions which arise from properties of the rectifiers and from the limitation to a specific voltage range. These restrictions enable one to establish a standard log characteristic which serves as a convenient basis for evaluation of experimental data.

The input voltage is limited to a given range, say $A \leqq E_i \leqq B$. The maximum output is restricted by the condition that the circuit does not amplify; i.e., the output voltage of the log circuit is never greater than that of the divider with the rectifier removed. Hence the log characteristic can at most be tangent to $E_0 = \frac{1}{2}E_i$ for equal series and load resistances and cannot cross this curve. If the log characteristic is to be tangent to $E_0 = \frac{1}{2}E_i$, the forward resistance of the rectifier must be infinite at the point of tangency and decrease in either direction. Since a resistance variation which involves a peak in mid-range has not been observed, it is assumed that the resistance is infinite at the lowest input voltage and decreases continuously for larger voltage.

A circuit having a logarithmic response satisfies

$$E_0 = K \ln E_i + C \tag{1}$$

where K and C are constants.

At $E_i = A$ it is assumed that $E_0 = \frac{1}{2}A$ and that the curve (1) is tangent to

$$E_0 = \tfrac{1}{2}E_i, \tag{2}$$

the output of the linear voltage divider corresponding to the circuit of Fig. 1.

Thus, K and C of (1) are determined. This gives for the standard response:

$$E_0 = \tfrac{1}{2}A\left(1 + \ln\frac{E_i}{A}\right). \tag{3}$$

Equation (3) represents the maximum output of a logarithmic voltage divider under the foregoing assumption. This can be shown by examining the derivative of the rectifier resistance Z with respect to the input voltage E_i. One obtains

$$Z' = R(K - E_0)/(E_i - 2E_0)^2 \tag{4}$$

where relation (1) and the resistance-voltage relationship for the voltage divider of Fig. 1 have been used ($R_S = R_L = R$). It is evident from (4) that, for Z' to be negative, E_0 must be greater than K. Hence, K must be equal to or less than $A/2$, the lowest output voltage.

Thus, under these conditions, the maximum possible output voltage from a divider which is perfectly logarithmic for input voltages greater than A is given by the standard characteristic, (3).

The use of a "biasing" voltage occurs in certain circuits. In most cases, this consists of placing a constant voltage, in series with the input signal, and a second, bucking voltage in the output to cancel the no-signal input current. It has been our experience that biasing arrangements of this type do not appreciably improve the logarithmic response.

Temperature Compensation

All semiconductors are temperature sensitive and invariably have a negative temperature coefficient of resistance. Certain nonlinear elements, such as thermistors, employ this attribute to advantage; in other cases, such as the application under consideration, this thermal sensitivity must be minimized.

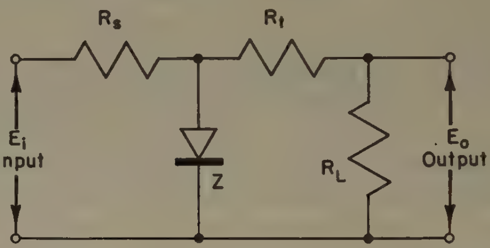

Fig. 2—Temperature-compensated logarithmic circuit.

Practical considerations lead immediately to the termperature-compensating logarithmic circuit such as shown in Fig. 2, employing a negative temperature coefficient resistance. For a given input voltage, it is desired that the output voltage remain constant with variations of temperature. Qualitatively, as the resistance of the rectifier decreases with increasing temperature, the resistance of the negative temperature coefficient (NTC) resistor should decrease in such a manner that the output voltage remains constant.

The best compensation is obtained by finding the lowest resistance NTC unit which overcompensates the output. This serves to minimize the insertion loss. The NTC unit is then shunted by a conventional variable resistance which is subsequently adjusted for best compensation. This process is illustrated by Figs. 3, 4, and 5 (opposite page).

Fig. 3 gives the uncompensated response of a circuit utilizing a copper oxide rectifier. This unit has a logarithmic output for input voltages above two volts. Fig. 4 gives the overcompensated response of the same rectifier, when an appropriate NTC resistor is placed in the circuit. This NTC unit was obtained from Keystone Carbon Company, type LE-701, 0.56 megohms at 100 degrees F. The temperature coefficient of resistance of this unit varies from -0.020/degree F. at 30 degrees F. to -0.016/degree F. at 100 degrees F.

Fig. 5 illustrates the best compensation obtainable with this circuit. In this instance, a conventional resistance of 3.3 megohms was found to be the optimum choice.

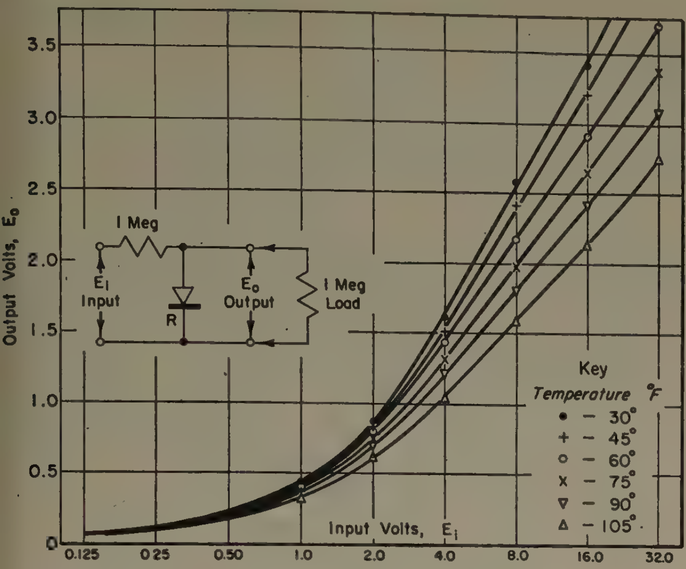

Fig. 3—Logarithmic characteristics of Bradley model CX18CA15H copper-oxide rectifier at various temperatures.

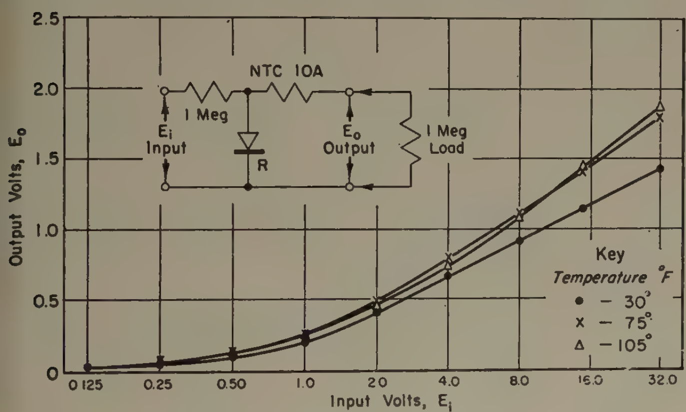

Fig. 4—Logarithmic characteristic of overcompensated Bradley rectifier.

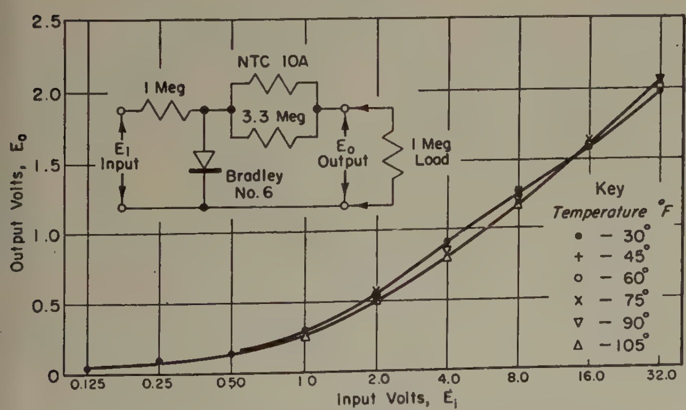

Fig. 5—Logarithmic characteristic of compensated Bradley rectifier CX18CA15H.

A measure of the temperature dependence is obtained by dividing the difference in the highest and lowest output voltage by the average output, all at a given input voltage level. Thus, along the usable portion of the characteristic, the spread at four volts input is 37 per cent of the average output in the uncompensated case. For the compensated circuit, the comparable figure is 12 per cent. Similarly, at 32 volts input, the uncompensated circuit yields 45 per cent; compensated, 3 per cent. Two other effects resulting from temperature compensation should be noted: (1) the output level is reduced, and (2) the range of logarithmic output has been extended down to about 1.0 volt input.

Fig. 6 is another illustration of temperature compensation by the same method. The upper curves give typical uncompensated response of a circuit utilizing a selenium rectifier. The lower curves give compensated response of the same circuit with indicated addition of an NTC resistor[4] shunted by 3 megohms.

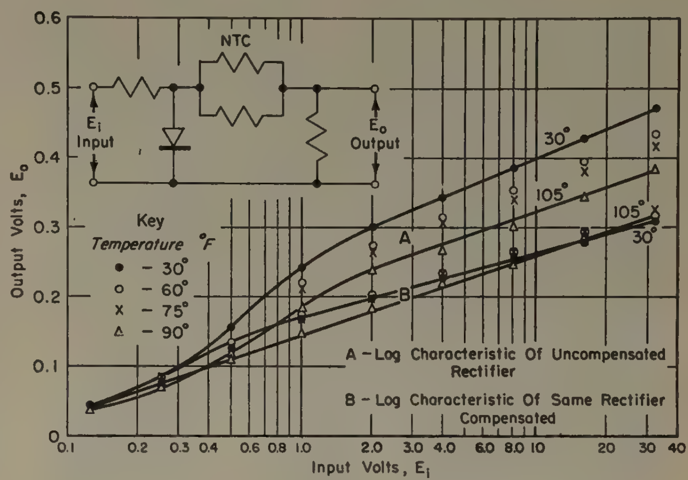

Fig. 6—Temperature compensation of GE 6RS5K84 rectifier.

It should be noted that exact temperature compensation requires a compensating circuit which is dependent on the *voltage input* as well as the temperature, since the temperature variation of resistance of these rectifiers also depends on the input level. For this reason, some attempts have been made to employ a compensating circuit in which the negative temperature coefficient resistor is replaced by a rectifier similar to the one used in the logarithm-sensitive portion of the circuit. However, these attempts were not as successful as those mentioned above.

Other Measurements

Some studies of the temperature and voltage stability of one particular rectifier were made. One unit was cycled more or less continuously between the temperature limits of 30 degrees and 105 degrees F. over a period of about two months. The characteristic of this unit was measured during each temperature cycle. No appreciable change was noted in its characteristics. Another unit was placed in the voltage divider circuit to which 8 volts dc were applied continuously for a period of 90 days. The response of this unit remained substantially unchanged during this time. In general, observations on a variety of germanium, selenium, and copper-oxide rectifiers indicate good temperature and

[4] Type LE-701, 0.15 megohm at 100 degrees F.

voltage stability if the application of the unit requires it to pass small currents in the forward direction only.

Since for dc input voltages the 6RS5K84 single disc selenium rectifier was found to exhibit logarithmic characteristics superior to any other type observed, measurements were made on several hundred of these units. The average response is shown in Fig. 7. The standard response has been included for comparison. The average characteristic is accurately logarithmic except at the lowest input voltage. The temperature compensation of this unit has already been mentioned.

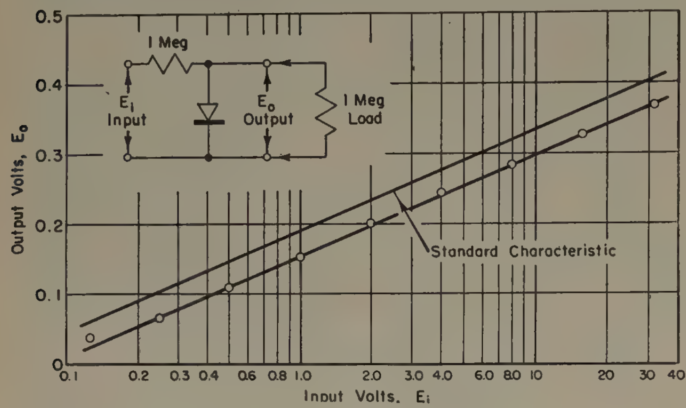

Fig. 7—Average dc logarithmic characteristic of 100 GE 6RS5K84 selenium rectifiers.

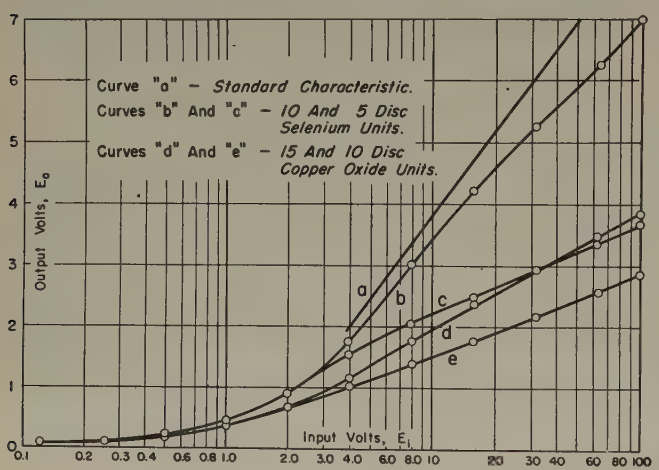

Fig. 8—Log characteristics of multidisc rectifiers.

In addition, data have been obtained on a variety of other single and multiple disc selenium and copper-oxide rectifiers. The characteristics of several multiple disc units are shown in Fig. 8. The high forward resistance of these units (compared to the series and load resistances of one megohm) makes them unsuitable as logarithmic elements for input voltages below approximately four volts. However, the response from 4 to 100 volts input is reasonably good. The standard response for $A=4$ volts is also indicated in Fig. 8. The output of these units increases with the number of discs in series. However, the linearity becomes poorer for units having more discs, although all units are logarithmic over a restricted range. Apparently because of very low-current densities, characteristics of these multiple disc rectifiers do not influence the circuit as nonlinear elements for input voltages up to 4 volts input.

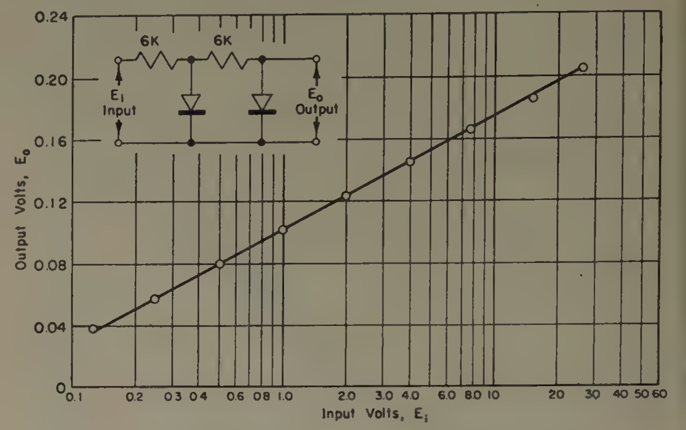

Fig. 9—Logarithmic characteristic of germanium rectifier, type IN65.

Early measurements were carried out on germanium crystal rectifiers of the IN34 type. The results indicated that these rectifiers are unsuitable as logarithmic elements in high resistance circuits, a fact which appears to be due in some way to their extremely low forward resistance as compared to the one megohm source resistance. The type IN65, a welded-contact diode, has substantially the same characteristic, and as Fig. 9 shows, a good logarithmic characteristic can be obtained with a combination of germanium diodes when much lower source impedances are used. This network is quite similar to that used in a logarithmic device produced commercially. Connecting diodes in parallel "front to back" in such a circuit permits its use with ac inputs at moderately high frequencies.

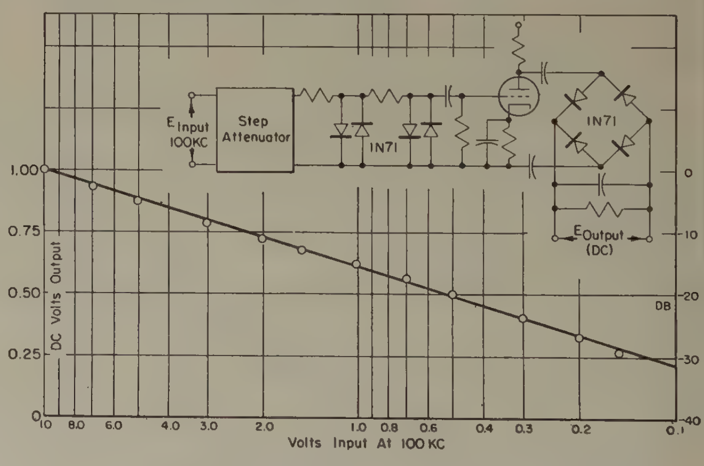

Fig. 10—Logarithmic detector and its response. Note direction of increasing input voltage.

A logarithmic detector circuit which accepts ac inputs at frequencies up to several hundred kc and gives a dc output proportional to the logarithm of the ac input voltage is shown in Fig. 10. The response of this circuit is also shown. No provision has been made for temperature compensation and the output is reasonably independent of ambient temperature variations once the unit has reached a uniform operating temperature.

Vacuum-Tube Detector and Converter for Microwaves Using Large Electron Transit Angles*

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Summary—**The following paper describes experimental and theoretical studies on a new phenomenon of microwave detection and conversion. Vacuum-tube diodes are used, but operating with very large electron-transit angles. It is shown that the cumulative energy received from the microwave field causes electrons to move out of the space charge region more rapidly than under dc conditions. This accounts for the detection and conversion action of the diode when a microwave signal is radiated into the interelectrode space.**

THIS PAPER DESCRIBES a new type of detection and conversion phenomenon which takes place in space-charge limited tubes operating with very large electron transit angles. This type of detection and conversion occurs in both vacuum tubes and ionized gas tubes. Because of noise considerations, the vacuum-tube operation is preferable. The behavior of ionized gases as detectors of microwave signals has been described in previous papers, (Ref. 1 and 2, p. 1123) hence this paper will deal only with vacuum-tube detectors and converters.

The transit angles involved in these tubes correspond to several hundred to thousands of cycles of the ac wave. In vacuum-tube operation, it has been customary to minimize electron-transit time. Analytical treatments have been concerned at most with transit angles up to several Pi radians. The large transit time operation in these tubes therefore involves entirely new considerations in tube design and analysis.

It is well known that conventional vacuum-tube detectors and converters experience declining sensitivities as the frequency of the impressed signal is increased into the ultra-high frequency or microwave region. One of the principal limitations at these frequencies is the electron-transit time. When the transit time becomes comparable to half the period of the microwave signal, the detection or conversion sensitivity vanishes. However, under certain conditions of operation, if a space-charge limited tube has very large transit angles, the detection and conversion action resumes, but operating on a wholly different principle. This very large transit-angle operation has not been described previously in the technical literature.

The development at present is still in the early stage, but preliminary experiments on cylindrical-diode vacuum tubes have yielded noise levels and minimum detectable signals at $\lambda = 3$ cm of the order of 13 db poorer than those of good 1N23 crystals. It is believed that substantial improvements can be effected by continued research.

Fig. 1 shows a vacuum-tube detector adapted to a crystal mount.

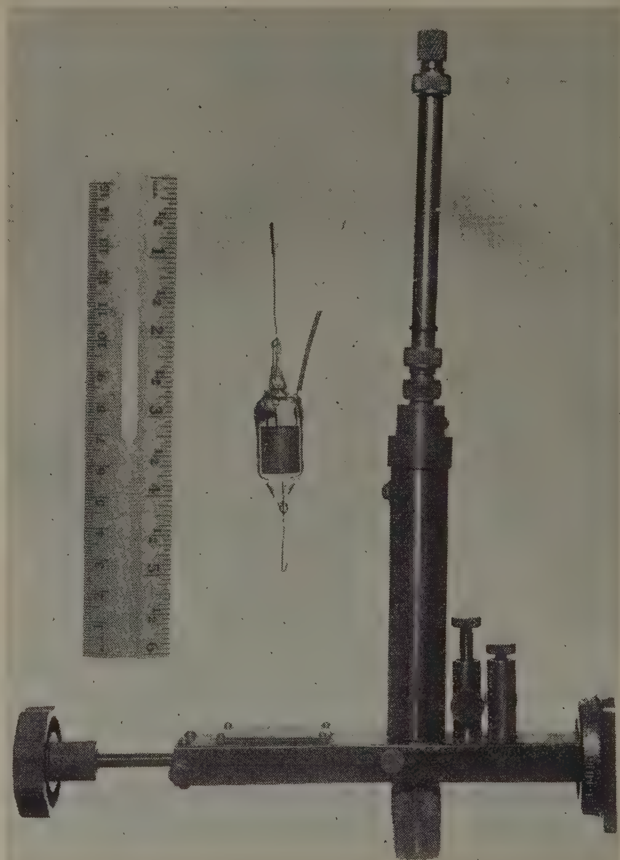

Fig. 1

If the performance of this type of detector-converter can be made comparable to that of crystal detectors, it would have many attractive advantages. For example, the tube is self-limiting at large microwave powers and burnouts such as frequently occur in crystal detectors cannot occur. This would make it possible to eliminate the costly TR box and associated equipment in radar receivers, since the sole purpose of such devices is to protect the crystal. Likewise, the tube is less susceptible to damage caused by mechanical shock, and a higher degree of uniformity in operating characteristics can be achieved in production quantities than is possible with crystals.

* Decimal classification: R134×R335×R339.2. Original manuscript received by the Institute, October 13, 1953; revised manuscript received, March 1, 1954.

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Fig. 2 shows the output of a tungsten-filament-diode detector with several different supply voltages, operating into a 100,000 ohm, resistance load, as compared with a good 1N23 crystal. Although the curves show supply voltages as high as 50v, it should be pointed out that in each case, the voltage from cathode to anode in the tube was at most a few volts positive.

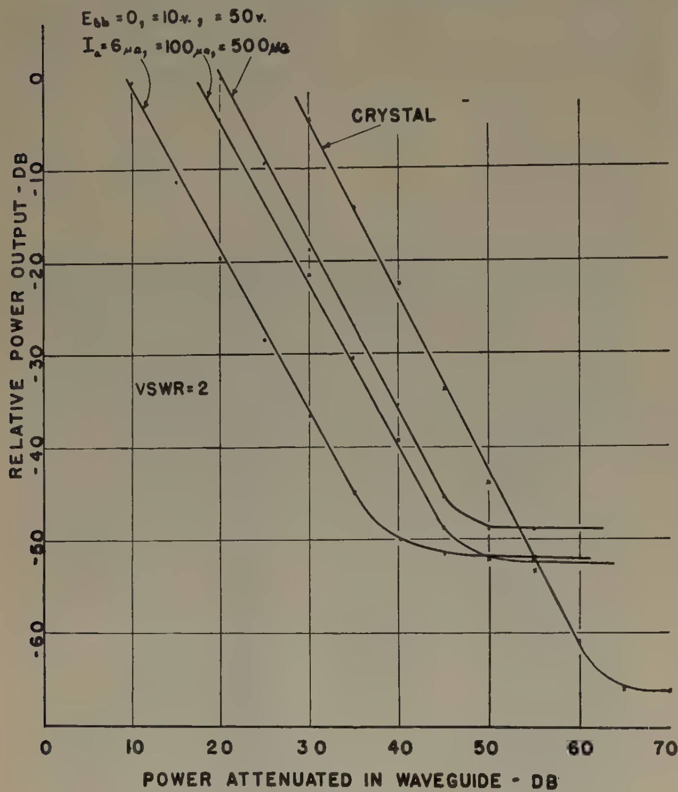


Fig. 2—Tungsten-filament diode compared with 1N23 crystal as a detector.

Fig. 3 shows the increment of dc plate current resulting from impressing the microwave field upon the diode, for various anode voltages. This shows maximum detection sensitivity at 2v for this particular tube.

The analytical behavior of electrons in vacuum tubes at high frequencies has been extensively treated by W. E. Benham, (Ref. 2, p. 1123) F. B. Llewellyn, (Ref. 3, p. 1123) and J. Muller (Ref. 4, p. 1123). Their methods, however, are based upon certain assumptions which are not valid for the large transit-angle tube operation described here and therefore it seems apparent that the functioning of this type of detector-converter could not be predicted from their analytical derivations. Specifically, their derivations assume: (a) all electrons have identical emission velocities, (b) there is a uniflow of electrons in one direction only, and (c) there is a one-to-one correspondence between the behavior of all electrons at a given space-time position and the electric field. None of these conditions exist in the vacuum tube detector-converter described here. It is believed that the detection-conversion phenomenon cannot be explained without considering bi-directional flow of electrons having a spectrum of emission velocities. Also, the previous derivations cannot be relied upon with any degree of validity at transit angles as large as those experienced in this large transit-angle detector. N. A. Begovich (Ref. 5) and G. Diemer (Ref. 6) used somewhat different approaches in considering the loading effect of a diode used as a resonant cavity, but the electron transit angles they considered were small, and their linear potential-distribution assumption does not fit our case.

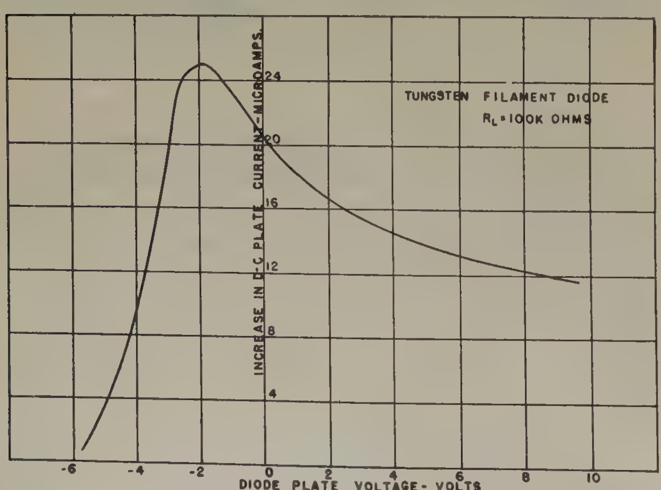


Fig. 3—The increment of dc plate current resulting from impressing the microwave field upon the diode.

In order to describe the behavior of this detector, consider the parallel plane diode of Fig. 4, which has a substantial space charge, producing a potential minimum.

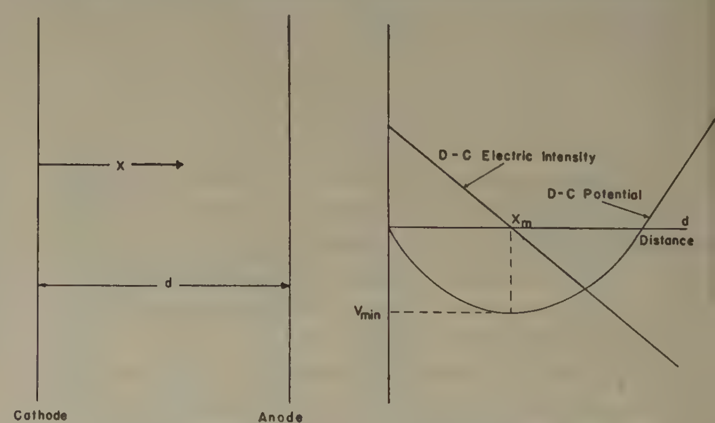


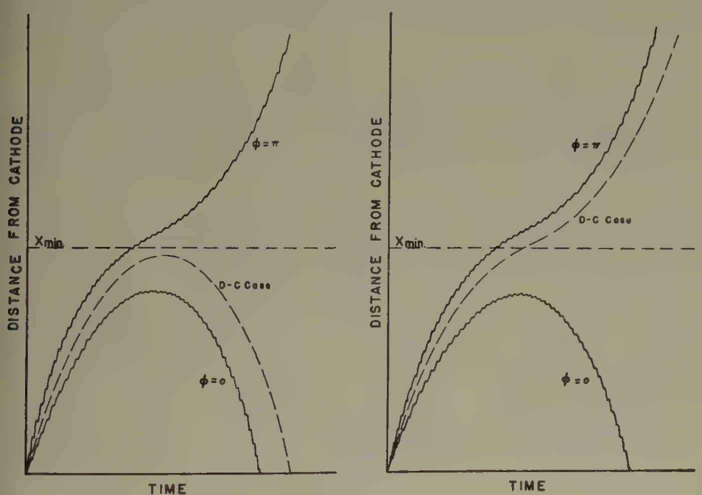
Fig. 4—The assumed dc potential and electric-intensity distributions.

When the microwave field is superimposed upon the dc field, this produces: (1) a rapid oscillatory motion superimposed upon the dc drift velocity, and (2) a cumulative acceleration or deceleration of the electron, depending upon the phase of the ac wave when the electron is emitted from the cathode. This effect is shown in Fig. 5, the dashed line representing the dc space diagram and the lines with superimposed oscillations, representing the electron travel in combined dc and ac fields.

The net effect of impressing the ac (microwave) field upon the dc field is to cause a reduction in average transit time of the electron. This in turn reduces the

space-charge density, raises the value of potential at the potential minimum and causes an increase in dc-plate current. This increase in plate current when microwave fields are impressed has been experimentally verified.

Consider four separate electrons. If only a dc field exists, two of the electrons would return to the cathode as shown by the single dashed line of Fig. 5(a) and two would travel to the anode, as shown by the single dashed line in Fig. 5(b). Now assume that a microwave field is superimposed upon the dc field. It will be further assumed that one of the two electrons of Fig. 5(a) is emitted from the cathode in such a phase with respect to the ac field that it experiences cumulative acceleration from the ac field. This causes the electron to pass through the potential minimum and advance to the plate. The other electron is emitted during the retarding phase of the ac field and receives cumulative deceleration, returning to the cathode earlier than it would if traveling in a purely dc field. Of course, there are other electrons which are emitted during such phases that they receive less cumulative acceleration or deceleration, hence their motion may be more nearly the dc drift motion, with a small cumulative ac oscillation.



(a) (b)

Fig. 5—The dc and ac space time diagrams.

Now consider two electrons in Fig. 5(b) which would under dc conditions travel to the plate. One is emitted in such a phase as to receive cumulative acceleration from the ac field, whereas the other receives cumulative deceleration. The first attains a velocity higher than the dc velocity and travels to the anode sooner than it would in a dc field. The second never passes through the potential barrier, but returns to the cathode.

The behavior of these four electrons can be summarized as follows:

Dc Operation	Dc and Ac Operation
a. electron returned to cathode	electron travels to anode
b. electron returned to cathode	electron returns to cathode *in less time than under dc operation*
c. electron traveled to anode	electron travels to anode *in less time than under dc operation*
d. electron traveled to anode	electron returns to cathode.

Now consider in a qualitative way the change which these four electrons would produce in space charge density as a result of impressing the ac (microwave) field. Electrons (2) and (3) both move out of the space charge region *faster* than they did under dc conditions. Electrons (1) and (4), taken together, act in opposite directions, hence they do not alter the space-charge density. The net effect of all four electrons is to *reduce* the space-charge density. But a reduction of space-charge density is accompanied by a slight rise in the potential curve. This in turn allows more electrons to flow to the anode, which accounts for the increase in anode current when the microwave field is impressed on the diode.

When a modulated-microwave signal is impressed upon the diode, the anode current rises and falls with the modulation amplitude, resulting in a detection of the signal. When two signals having slightly different frequencies are impressed, the resultant ac field rises and falls at a frequency equal to the difference frequency, hence there is an output at the difference frequency, which explains the converter action. The only possible limitation as a converter seems to be that the electron-transit time from cathode to anode must be small in comparison with the period of the difference frequency. But this condition is satisfied at all practical intermediate frequencies.

The following is an analytical approach of a parallel-plane diode which accounts for the detection-conversion phenomenon, although it is based upon certain approximations of potential distribution.

Equation of Motion

Consider the diode of Fig. 4. Let x be the distance of electron travel in time t and e/m be the charge to mass ratio of an electron. The dc electric intensity is $E_0(x)$ and the ac intensity $E_1(x)\sin(\omega t+\phi)$. An electron is assumed to be emitted from the cathode at $t=0$, when the ac intensity has a phase angle ϕ. Newton's law of motion is then

$$\frac{d^2x}{dt^2} = -\frac{e}{m}\left[E_0(x) + E_1(x)\sin(\omega t + \phi)\right]. \qquad (1)$$

Since best detection-conversion characteristics are obtained when very low plate voltages are applied, the dc-potential distribution curve has a pronounced potential minimum. It is therefore assumed that the dc potential curve is parabolic, given by

$$V_0(x) = \frac{2V_m}{x_m^2}\left(\frac{x^2}{2} - xx_m\right) \qquad (2)$$

with a corresponding dc electric intensity

$$E_0(x) = -\frac{dV_0(x)}{dx} = -\frac{2V_m}{x_m^2}(x - x_m). \tag{3}$$

It is not necessary to have an exact determination of the ac intensity distribution, since it is believed that the detection-conversion phenomenon would take place for almost any arbitrary ac-field distribution. We assume that the ac field is uniform, hence $E_1(x)$ is constant. Let

$$a_1^2 = \frac{e}{m}E_1 \tag{4}$$

where a_1 is a constant. Also let

$$a_0^2 = \frac{2eV_m}{mx_m^2}. \tag{5}$$

Then (1) becomes

$$\frac{d^2x}{dt^2} = a_0^2(x - x_m) - a_1^2 \sin(\omega t + \phi) \tag{6}$$

where a_1 and a_0 have different dimensions. The successive integration of (6), with the substitution of the initial conditions $v = v_e$ and $x = 0$ when $t = 0$, yields the expressions for the electron velocity and displacement:

$$v = (v_e - \omega k \cos \phi)\cosh a_0 t - a_0(x_m + k \sin \phi)\sinh a_0 t + \omega k \cos(\omega t + \phi) \tag{7}$$

$$x = x_m + \left(\frac{v_e}{a_0} - \frac{\omega k}{a_0}\cos\phi\right)\sinh a_0 t - (x_m + k\sin\phi)\cosh a_0 t + k \sin(\omega t + \phi) \tag{8}$$

where

$$k = \frac{a_1^2}{a_0^2 + \omega^2}.$$

The dc velocity and displacement are obtained by setting $k=0$ in (7) and (8),

$$v_0 = v_e \cosh a_0 t - a_0 x_m \sinh a_0 t \tag{9}$$

$$x = x_m + \frac{v_e}{a_0}\sinh a_0 t - x_m \cosh a_0 t. \tag{10}$$

The terms in (7) and (8) multiplied by $\cosh a_0 t$ and $\sinh a_0 t$ represent the functions which either increase or decrease cumulatively as time increases. The terms $\sin(\omega t + \phi)$ and $\cos(\omega t + \phi)$ are purely oscillatory. It is these terms which produce the wiggle on the curves of Fig. 5. Since the electron-transit angles are extremely large, it is assumed that these latter terms are small in comparison with the nonoscillatory terms for most of the electrons. As a reasonable approximation, they will be discarded in the subsequent analysis, i.e., we shall consider only the nonoscillatory effects of the ac field.

By combining (7) and (8), the oscillatory terms being neglected, the variable t can be eliminated and the expression for electron-velocity v can be written in terms of x as

$$v = +\sqrt{(v_e - \omega k \cos\phi)^2 - a_0^2(x_m + k\sin\phi)^2 + a_0^2(x_m - x)^2} \quad \text{for } x \geqq x_m$$

or

$$v = \pm\sqrt{(v_e - \omega k \cos\phi)^2 - a_0^2(x_m + k\sin\phi)^2 + a_0^2(x_m - x)^2} \quad \text{for } x < x_m \tag{11}$$

where the sign will depend upon whether the electron is going toward the anode or returning to the cathode. These expressions are not accurate for x very close to the cathode because of the neglect of oscillatory terms in (7) and (8).

The velocity in the dc case may be obtained either by combining (9) and (10), or by merely setting $k=0$ in (11),

$$v_0 = +\sqrt{v_e^2 - (a_0 x_m)^2 + a_0^2(x_m - x)^2} \quad \text{for } x \geqq x_m$$

or

$$v_0 = \pm\sqrt{v_e^2 - (a_0 x_m)^2 + a_0^2(x_m - x)^2} \quad \text{for } x < x_m \tag{12}$$

where the sign again is determined by the traveling direction of electrons.

Initial Velocity and Phase Angle

To find the lowest initial velocity with which an electron will get through the potential minimum, we may let $x = x_m$ and $t = T_m$ in (8), yielding

$$\tanh a_0 T_m = \frac{a_0(x_m + k\sin\phi)}{v_e - \omega k \cos\phi}. \tag{13}$$

Since the hyperbolic tangent of a real angle has an asymptotic value of unity, when the right side of (13) is greater than unity, T_m is imaginary. For this condition, the electron will never get to the potential minimum x_m, but will return to the cathode. Thus an electron will pass x_m only if

$$\tanh a_0 T_m = \frac{a_0(x_m + k\sin\phi)}{v_e - \omega k \cos\phi} < 1. \tag{14}$$

It will return to the cathode if

$$\tanh a_0 T_m = \frac{a_0(x_m + k\sin\phi)}{v_e - \omega k \cos\phi} > 1. \tag{15}$$

These equations show that the emission velocity for electrons to reach the anode is

$$v_e > \omega k \cos\phi + a_0(x_m + k\sin\phi) \tag{16}$$

in the ac case; and

$$v_e > a_0 x_m \tag{17}$$

in the dc case.

In (16) the minimum value of v_e for electrons to reach the anode is dependent upon ϕ. There is a range of values of v_e for which the electrons either do or do not

pass through the potential minimum, depending upon the value of ϕ. Within this range, we can solve for the critical values of ϕ at which electrons barely pass through the potential minimum by letting (14) equal unity. This yields

$$\begin{aligned}
\sin\phi_1 &= \frac{a_0(v_e - a_0x_m) + \omega\sqrt{k^2(\omega^2+a_0^2)-(v_e-a_0x_m)^2}}{k(\omega^2+a_0^2)} \\
\cos\phi_1 &= \frac{\omega(v_e - a_0x_m) - a_0\sqrt{k^2(\omega^2+a_0^2)-(v_e-a_0x_m)^2}}{k(\omega^2+a_0^2)} \\
\sin\phi_2 &= \frac{a_0(v_e - a_0x_m) - \omega\sqrt{k^2(\omega^2+a_0^2)-(v_e-a_0x_m)^2}}{k(\omega^2+a_0^2)} \\
\cos\phi_2 &= \frac{\omega(v_e - a_0x_m) + a_0\sqrt{k^2(\omega^2+a_0^2)-(v_e-a_0x_m)^2}}{k(\omega^2+a_0^2)}.
\end{aligned} \tag{18}$$

In studying this set of equations, we see that, if $v_e = a_0x_m \pm k\sqrt{\omega^2+a_0^2}$, the term $\sqrt{k^2(\omega^2+a_0^2)-(v_e-a_0x_m)^2}$ vanishes, and it yields $\phi_1=\phi_2$. This leads to the following conclusion:

(a) The electrons having $v_e > a_0x_m + k\sqrt{\omega^2+a_0^2}$ will get through the potential minimum point regardless of the phase angle ϕ at which they are emitted from the cathode.

(b) The electrons having $v_e < a_0x_m - k\sqrt{\omega^2+a_0^2}$ will never reach the potential minimum point regardless of the value of ϕ.

(c) The electrons having v_e in the range $a_0x_m + k\sqrt{\omega^2+a_0^2}$ to $a_0x_m - k\sqrt{\omega^2+a_0^2}$ will either get through the potential minimum point, or not, depending upon the values of ϕ. This is shown in Fig. 6 in which the limits of ϕ are determined by (18).

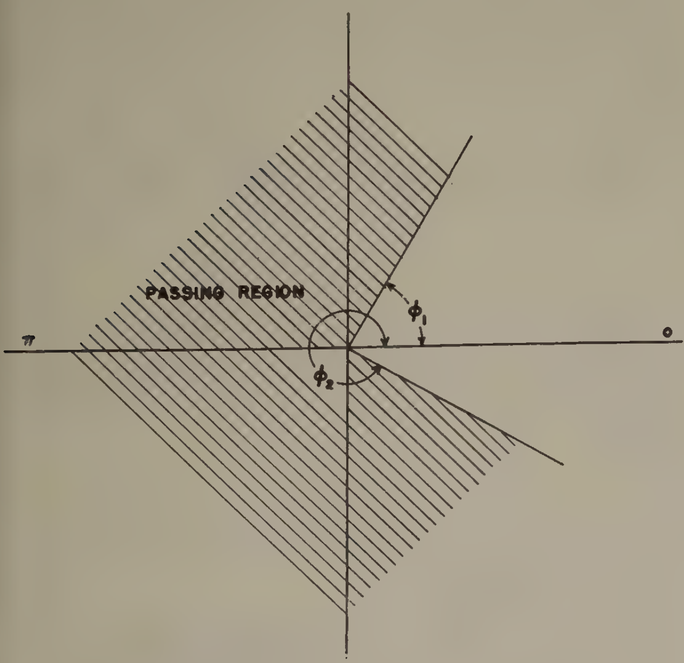


Fig. 6—Electrons pass through potential minimum for $\phi_1 < \phi < \phi_2$ as given by (18)

The electrons emitted by a cathode have very nearly Maxwell-Boltzman velocity distribution as given by

$$dN = \frac{m}{KT} N v_e e^{-\frac{m}{2KT}v_e^2} dv_e \tag{19}$$

and

$$N = \int_0^\infty \frac{m}{KT} N v_e e^{-\frac{m}{2KT}v_e^2} dv_e \tag{20}$$

where

v_e = normal component of emission velocity
m = electron mass
d_N = number of emitted electrons per unit time per unit area in the velocity range v_e to v_e+dv_e
N = total number of emitted electrons per unit time per unit area
K = Boltzman's constant
T = absolute temperature.

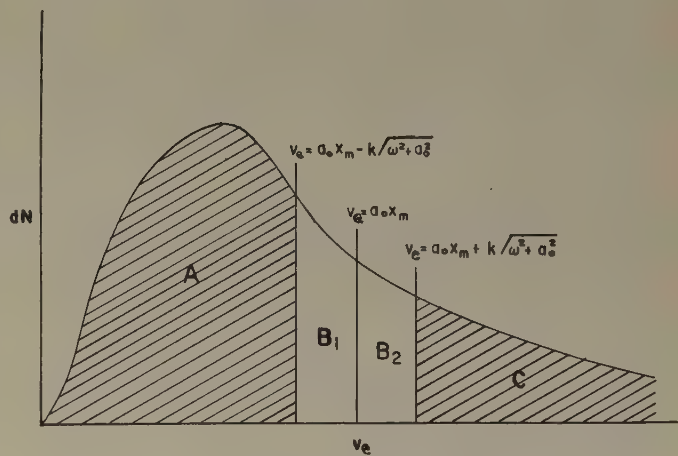


Fig. 7—The velocity-distribution diagram..

Fig. 7 shows the different situations encountered in a pure dc case and a superimposed dc and ac case. In the dc case, the electrons of regions C and B_2 have sufficient emission velocities to pass through the potential minimum while those of regions B_1 and A do not. When the ac field is impressed, all electrons in region A will return to the cathode, and those of region C will reach the anode, regardless of the value of ϕ. But, the electrons of parts B_1 and B_2 will either get through x_m, or not, depending upon the ϕ-value as discussed in (c).

Convection Currents

The total convection current at any position x may be expressed in terms of the current distribution function $i(x, v, t)dv$, which is defined as the current at the position x at time t due to electrons having velocities in the range v to $v+dv$, as

$$I(x, t) = \int_{v_l}^\infty i(x, v, t)dv \tag{21}$$

where v_l = lowest electron velocity at x at time t.

The Liouville's theorem (Ref. 7, p. 1123) states that

$$\frac{i(x, v, t)}{v} = \frac{i(x', v', t')}{v'} \tag{22}$$

where the primed quantities represent a new space-time position.

The current distribution function at the cathode may be written as

$$i(o, v_e)dv_e = I_s \frac{m}{KT} v_e e^{-\frac{m}{2KT} v_e^2} dv_e. \tag{23}$$

Then, according to (22) and (21), we have

$$i(x, v)dv = I_s \frac{m}{KT} ve^{-\frac{m}{2KT} v_e^2} dv \tag{24}$$

and

$$I(x, v) = I_s \frac{m}{KT} \int_{v_l}^{\infty} ve^{-\frac{m}{2KT} v_e^2} dv \tag{25}$$

where $I_s = Ne$ is the saturation current.

(a) *Dc case:* The velocity expression (12) may be transformed into

$$v_e^2 = v_0^2 + (a_0 x_m)^2 - a_0^2 (x_m - x)^2. \tag{26}$$

For $x \geqq x_m$, the lowest electron velocity at x is

$$v_l = a_0 (x - x_m) \tag{27}$$

as obtained by letting $v_e = a_0 x_m$ in (12).

Substituting (26) and (27) into (25), we have the total convection current at x

$$I_{do}(x) = I_s e^{-\frac{m}{2KT}(a_0 x_m)^2} \tag{28}$$

which is independent of x as expected.

(b) *Ac case:* By a similar approach, (11) may be transformed into

$$v_e^2 = \omega k \cos \phi + \sqrt{v^2 + a_0^2 (x_m + k \sin \phi)^2 - a_0^2 (x_m - x)^2} \tag{29}$$

and, for $x \geqq x_m$, the lowest velocity at x is

$$v_l = a_0 (x - x_m) \tag{30}$$

as obtained by letting $v_e = \omega k \cos \phi + a_0 (x_m + k \sin \phi)$ in (11).

Substituting (29) and (30) into (25), we have

$$I(x) = I_s \frac{m}{KT} \int_{a_0(x - x_m)}^{\infty} ve^{-\frac{m}{2KT}\left[\omega k \cos \phi + \sqrt{v^2 + a_0^2 (x_m + k \sin \phi)^2 - a_0^2 (x_m - x)^2}\right]^2} dv. \tag{31}$$

The convection current $\overline{I(x)}$ averaged over ϕ will be

$$\overline{I(x)} = \frac{1}{2\pi} I_s \frac{m}{KT} \int_0^{2\pi} \int_{a_0(x - x_m)}^{\infty} ve^{-\frac{m}{2KT}\left[\omega k \cos \phi + \sqrt{v^2 + a_0^2 (x_m + k \sin \phi)^2 - a_0^2 (x_m - x)^2}\right]^2} dv d\phi. \tag{32}$$

If we let $x = x_m$, we shall have the average convection current at the potential minimum

$$\overline{I(x_m)} = \frac{1}{2\pi} I_s \frac{m}{KT} \int_0^{2\pi} \int_0^{\infty} ve^{-\frac{m}{2KT}\left[\omega k \cos \phi + \sqrt{v^2 + a_0^2 (x_m + k \sin \phi)^2}\right]^2} dv d\phi. \tag{33}$$

Eq. (32) expresses the average convection current at position x. The average current at the cathode can be expressed in a simpler way. By substituting (16) and (23) into (25), the current at the cathode becomes

$$I(o) = I_s \frac{m}{KT} \int_{\omega k \cos \phi + a_0 (x_m + k \sin \phi)}^{\infty} v_e e^{-\frac{m}{2KT} v_e^2} dv_e$$
$$= I_s e^{-\frac{m}{2KT}\left[\omega k \cos \phi + a_0 (x_m + k \sin \phi)\right]^2}. \tag{34}$$

Averaging this over ϕ, we have

$$\overline{I(o)} = \frac{1}{2\pi} I_s \int_0^{2\pi} e^{-\frac{m}{2KT}\left[\omega k \cos \phi + a_0 (x_m + k \sin \phi)\right]^2} d\phi. \tag{35}$$

Then, the ratio of (35) to (28) will indicate the amount of increment of plate current resulting from impressing the microwave field upon the dc field. This is

$$\frac{\overline{I(o)}}{I_{d.c.}} = \frac{1}{2\pi} \int_0^{2\pi} e^{-\frac{m}{2KT}(\omega k \cos \phi + a_0 k \sin \phi)(\omega k \cos \phi + a_0 k \sin \phi + 2 a_0 x_m)} d\phi. \tag{36}$$

However, if some reasonable approximations are made, the current expression can be expressed in a more explicit form.

By a simple transformation, the following equality is obtained:

$$\omega k \cos \phi + a_0 (x_m + k \sin \phi) = \sqrt{(\omega k)^2 + (a_0 k)^2} \cos (\phi + \theta) + a_0 x_m \tag{37}$$

where $\theta = \tan^{-1}(a_0 k / \omega k)$.

A thorough investigation of practical cases has shown that the values of both (ωk) and $(a_0 k)$ are much smaller than that of $(a_0 x_m)$, their ratios are not larger than one-

to-ten. Consequently, it is quite reasonable to write current (35), in combination with (37), as

$$I = I_s \frac{1}{2\pi}\int_0^{2\pi} e^{-\frac{m}{2KT}\left[(a_0x_m)^2 + 2a_0x_m\sqrt{(\omega k)^2+(a_0k)^2}\cos(\phi+\theta)\right]} d\phi$$

$$= I_s e^{-\frac{m}{2KT}(a_0x_m)^2}$$

$$\cdot \frac{1}{2\pi}\int_0^{2\pi} e^{-\frac{m}{KT}a_0x_m\sqrt{(\omega k)^2+(a_0k)^2}\cos(\phi+\theta)} d\phi$$

$$= I_s e^{-\frac{m}{2KT}(a_0x_m)^2} I_0\left(\frac{m}{KT}a_0x_m\sqrt{(\omega k)^2+(a_0k)^2}\right) \tag{38}$$

where

$$I_0\left(\frac{m}{KT}a_0x_m\sqrt{(\omega k)^2+(a_0k)^2}\right)$$

is the modified Bessel's function of the first kind of zero order, which is always larger than one.

By combining (38) and (28), we have

$$\frac{\bar{I}}{I_{dc}} = I_0\left(\frac{m}{KT}a_0x_m\sqrt{(\omega k)^2+(a_0k)^2}\right). \tag{39}$$

It should be noted that this ratio is always larger than unity which agrees with the experimental results.

The function

$$\left(\frac{m}{KT}a_0x_m\sqrt{(\omega k)^2+(a_0k)^2}\right)$$

is rather involved. It contains several parameters which are defined by (4), (5), and (8).

An analysis of power transferred from electric field to electrons, or vice versa, in both the dc case and the superposed ac and dc case, taking into the consideration of velocity distribution of electrons, has been worked out. This analysis, which is not included here, has shown that on the average electrons will receive more power from electric field in the latter case than in the former as a result of impressing the ac (microwave) field. This analytical result also accounts for the influence of the microwave field in causing the electrons to move out of the space charge region more rapidly.

Conclusion

Experimentally, both microwave detection and conversion phenomena have been observed to take place in space-charge limited cylindrical diodes operating with very large transit angles. The foregoing theoretical approach has dealt with parallel-plane diodes, rather than cylindrical ones. Expressions have been obtained for the electron velocity, in superposed dc and ac fields, which is valid for large electron transit times, based upon certain assumptions regarding potential distribution. The effect of Maxwell-Boltzman distribution of electron emission velocities has been taken into consideration. The net effect of the ac field is to remove the electrons from the space charge region (either to the anode or to cathode) more rapidly than would occur under dc conditions. This reduction in space charge density causes the potential curve to rise, thus allowing more electrons to pass through the potential minimum to the anode. This accounts for the observed increase in anode current when the microwave energy is impressed and also accounts for the detection and conversion phenomena.

Acknowledgment

The authors wish to acknowledge gratefully the fact that this project was made possible by a grant from the National Science Foundation.

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Discussion on

"The Detection of a Sine Wave in the Presence of Noise"*

T. G. SLATTERY

W. W. Peterson:[1] In the above article, the author makes the statement that the likelihood ratio λ on the boundary curve in his sample space should be equal to

$$\lambda = \frac{A}{1 - B} \tag{1}$$

where

$$A = \int P_s(X_1X_2)dX_1dX_2$$

left of boundary

$$B = \int P_n(X_1X_2)dX_1dX_2 \tag{2}$$

right of boundary.

This equation is not correct. Actually, if the region to the left of the boundary is the region for which one will say there is a signal present, then, by (4),

$$P_s(X_1X_2) - \lambda P_N(X_1X_2) > 0$$

in the entire region. Integrating this expression over that region leads to:

$$\lambda < \frac{A}{1 - B} \cdot \tag{3}$$

The other hypothesis leads to the reverse inequality sign.

A consequence of this error is that the concluding paragraph of section I of the paper is not true. For a given set of signals and noise the best detection filter is not specified by both the probability of detection required, and the false alarm probability which can be tolerated. *Rather it is completely specified when the false alarm probability which can be tolerated is given.* Then the probability of detection can be calculated.

Section II-B is misleading. It states that Shannon and Singleton have proved that linear filters are best only when the signal has a Gaussian amplitude distribution. Actually they showed that if the noise and the signal have Gaussian amplitude distributions, then the best filter is a linear filter. Furthermore, they were discussing filters for smoothing and prediction, whereas this article is concerned with signal detection filters.[2] Woodward and Davies[3-5] have shown that in any case where the signal is known exactly, and the noise is Gaussian, the best detection filter is a linear filter.

* T. G. Slattery, "The detection of a sine wave in the presence of noise by the use of a nonlinear filter," PROC. I.R.E., vol. 40, pp. 1232–1236; October, 1952.

[1] University of Michigan, Engineering Research Inst., Ann Arbor, Mich.

[2] The distinction is pointed out in L. A. Zadeh and J. R. Ragazzini, "Optimum filters for the detection of signals in noise," PROC. I.R.E., vol. 40, pp. 1223–1231; October, 1952.

[3] I. L. Davies, "On determining the presence of signals in noise," *Proc. I.E.E.* (London), vol. 99, pp. 45–51; March, 1952.

[4] P. M. Woodward and I. L. Davies, "Information theory and inverse probability in telecommunication," *Proc. I.E.E.* (London), vol. 99, pp. 37–44; March, 1952.

[5] P. M. Woodward, "Information theory and the design of radar receivers," PROC. I.R.E., vol. 39, pp. 1521–1524; December, 1951.

CORRECTION

E. R. Kretzmer, author of the paper, "An Amplitude Stabilized Transistor Oscillator," which appeared on pages 391–401, of the February, 1954 issue of the PROCEEDINGS OF THE I.R.E., has brought the following corrections to the attention of the editors:

In the summary preceding the article "break-up effect" should read "break-down effect."

In the first paragraph of page 394 "Zener action" should be changed to "regulator action."

The subscript z should be changed to r wherever it appears: Fig. 2—E_r; Definitions of Symbols—e_r, E_r; Fig. 8—e_r; Equation 20, the following sentence and footnote 8—R_r. In the Definitions of Symbols, R_N should read R_r.

(All of these errors resulted from a recent change in terminology: what was formerly referred to as "Zener effect," "Zener voltage," and "Zener diode," is now commonly called "break-down effect," "break-down or regulator voltage," and "regulator diode," respectively.)

In connection with the silicon junction diode (Fig. 3), the following reference should have been cited: G. L. Pearson and B. Sawyer, "Silicon p-n junction alloy diodes," PROC. I.R.E., vol. 40, pp. 1348–1351; November, 1952.

Two Network Theorems Concerning Change of Voltage Reference Terminal*

JACOB SHEKEL†, ASSOCIATE, IRE

Summary—**The first theorem describes the change in the admittance matrix corresponding to the change of the voltage reference terminal ("ground" terminal). The second theorem shows that the determinant of the admittance matrix is invariant under such transformations.**

A GENERAL THEOREM brought by Kron[1] states that if a network is represented by the admittance matrix $\mathbf{Y}$, and the voltages of the terminals are transformed from $\mathbf{v}$ to $\bar{\mathbf{v}}$ by

$$\bar{\mathbf{v}} = \mathbf{P}\mathbf{v}, \tag{1}$$

the admittance matrix undergoes a congruent transformation to

$$\bar{\mathbf{Y}} = \mathbf{P}'\mathbf{Y}\mathbf{P} \tag{2}$$

($\mathbf{P}'$ is the tranpose of $\mathbf{P}$).

Applying this general transformation to the following problem:

An $(n+1)$-terminal network is analyzed with all voltages referred to the $(n+1)$-th terminal, and represented by an $n\times n$ admittance matrix $\mathbf{Y}$. How is the admittance matrix transformed if the r-th $(r<n+1)$ terminal is to be used as a new voltage reference terminal, and the old reference terminal denoted by r?

The transformation amounts, first, to a change in terminal numbering from

$$1, 2, \cdots, r, \cdots, n, 0$$

to

$$1, 2, \cdots, 0, \cdots, n, r$$

respectively, where the reference terminal is numbered 0, and then a change of voltages. The "new" voltages $\bar{\mathbf{v}}$, referred to the "new" reference terminal, are related to the "old" voltages $\mathbf{v}$ by

$$\bar{v}_i = v_i - v_r \quad (i = 1, 2, \cdots, n; i \neq r) \tag{3}$$

and the "new" voltage of the "old" reference terminal is the negative of the "old" voltage of the "new" reference terminal,

$$\bar{v}_r = -v_r. \tag{4}$$

Equations (3) and (4) together correspond to a transformation matrix $\mathbf{P}$ which is obtained from the $n\times n$ unit matrix by changing the r-th column into a column of (-1)'s. To take a specific example: a 6-terminal network, described by a 5×5 admittance matrix, has the voltage reference terminal interchanged with terminal 4, then

$$\mathbf{P} = \begin{Vmatrix} 1 & 0 & 0 & -1 & 0 \\ 0 & 1 & 0 & -1 & 0 \\ 0 & 0 & 1 & -1 & 0 \\ 0 & 0 & 0 & -1 & 0 \\ 0 & 0 & 0 & -1 & 1 \end{Vmatrix}. \tag{5}$$

Noting how $\mathbf{P}$ may be obtained from a unit matrix by elementary column operations,[2] transformation (2) results in the following:

Theorem 1. The required transformation (in the problem formulated above) is effected by substituting the negative of the sum of all the columns of $\mathbf{Y}$ for the r-th column, and then substituting the negative sum of all the rows of the resulting matrix for the r-th row.

The determinant of $\mathbf{P}$ may be obtained through expansion by the elements of the r-th column. The cofactor of every term off the principal diagonal contains a column of zeros; the only nonzero co-factor is that of P_{rr}, and is equal to 1, so that

$$|\mathbf{P}| = -1$$

and, from (2),

$$\begin{aligned} |\bar{\mathbf{Y}}| &= |\mathbf{P}'| \times |\mathbf{Y}| \times |\mathbf{P}| \\ &= (-1) \times |\mathbf{Y}| \times (-1|-) \\ |\bar{\mathbf{Y}}| &= |\mathbf{Y}|. \end{aligned} \tag{6}$$

Theorem 2. The determinant of the admittance matrix of a network does not depend on the voltage reference terminal chosen for the representation.

* Decimal classification: R143. Original manuscript received by the IRE, August 17, 1953.

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[1] G. Kron, "Tensor Analysis of Networks," chap. IV, "The transformation tensor," pp. 102–104, John Wiley and Sons, Inc., New York, N. Y.; 1939. Kron's formulas are about impedance matrices and transformation of mesh currents, but the application to admittance matrices and node voltages is straightforward.

[2] S. Perlis, "Theory of Matrices," chap. 3, "Equivalence, rank and inverses," Addison-Wesley Press, Inc., Cambridge, Mass.; 1952.

Some Techniques for Network Synthesis*

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Summary—Part I: Synthesis of Immittances with Two Poles and Two Zeros. Cauer's continued-fraction technique is generalized for use in realizing RLC network immittances. Using this method, element values are determined by simple processes of "forward" and "reverse" division.

Immittances $F(p)$ with two poles and zeros may be grouped in three classes according to the frequency $j\omega_m$ at which $Re\ F(j\omega)$ is a minimum. For Class I, $\omega_m=0$; for Class II, $\omega_m=\infty$; and for Class III, ω_m is finite. Class I and II $F(p)$ can be realized immediately by making a continued-fraction expansion. If both $F(p)$ and $1/F(p)$ are Class III, to obtain a realization without unity-coupled coils the function must be split into two terms which may then be expanded in continued fractions. Simple formulas are presented which enable one to easily classify $F(p)$ and determine an appropriate realization method.

Part II: A Constant-Resistance Ladder for Transfer Function Synthesis. The physical factors that determine the poles and zeros of a transfer function are examined. By use of physical insight, a design procedure for a constant-resistance ladder network is arrived at. This ladder network is found to have the same realm of application as the conventional, RLC, constant-resistance bridged-T. However, the ladder has the advantages of: fewer elements, more flexibility, and requiring less flat loss if the bridged-T requires flat loss. Complicated transfer functions may be realized in a chain of ladder sections having arm immittances with only two poles and two zeros. The techniques described in Part I may be utilized to facilitate the design of such sections.

Part I: Synthesis of Immittances with Two Poles and Two Zeros

SYSTEMS HAVING transfer functions of any complexity can be synthesized in a chain of structures whose component impedances have no more than two poles and two zeros. Some structures which may be used in this manner are the constant-resistance lattice, the constant-resistance bridged-T, and a constant-resistance ladder structure which will be described in Part II of this paper. In Part I we shall endeavor to see how we can find the simplest realization for any given impedance or admittance with two poles and two zeros (minimum number of circuit elements required may vary from two to nine if unity-coupled coils are excluded). In Part II a constant-resistance ladder structure having advantages for the synthesis of transfer functions will be introduced. The techniques described in Part I will be seen to be of considerable help in the synthesis of such networks.

The Conditions for Realizability

Any impedance or admittance function having two poles[1] may be represented in the form

$$F(p) = \frac{gp^2 + ap + b}{p^2 + ep + d}. \tag{1}$$

Following Bode,[2] we shall call $F(p)$ an *immittance* since it may be regarded equally well as either an impedance or an admittance. A necessary and sufficient condition for $F(p)$ to be realizable as a physical, passive imittance is that $F(p)$ must be what Brune[3] has defined as a *positive-real function.*[4] It can be seen that in the case of the relatively simple function (1) the conditions for p-r character may be reduced to:

A. All nonzero coefficients must be real and positive.

B. $Re\ F(j\omega) \geqq 0$ for all ω.

Condition A is easily checked, but we shall wish to have a simple method for checking condition B.

Note that any immittance function can be expressed as

$$F(p) = \frac{m_1 + n_1}{m_2 + n_2}, \tag{2}$$

where the m's are polynomials with only even-powered terms and the n's are polynomials with only odd-powered terms. As Guillemin shows[5]

$$Re\,F(j\omega) = \left.\frac{m_1m_2 - n_1n_2}{m_2^2 - n_2^2}\right|_{p=j\omega}. \tag{3}$$

From (3), it can be seen that if p-r condition A is satisfied, then p-r condition B will be satisfied if the equation

$$m_1m_2 - n_1n_2 = 0 \tag{4}$$

has no roots of odd multiplicity on the $j\omega$ axis. This is so because the numerator of (3) will have odd-order zeros on the $j\omega$ axis if $Re\ F(j\omega)$ swings from positive values to negative values. Constructing (4) from (1) and using some elementary theory of equation it can be seen that (4) will have no roots of odd multiplicity on the $j\omega$ axis if

$$b - ae + dg \leqq 2\sqrt{gbd}. \tag{5}$$

Thus we arrive at the important conclusion that if p-r condition A *is satisfied for* (1), *then condition* B *will be satisfied if, and only if* (5) *is satisfied.*

Three Classes of Functions

The nature of the $Re\ F(j\omega)$ characteristic can tell us much about how difficult a given immittance will be to realize. In general we can classify immittances of the form of (1) in three classes: Class I having $Re\ F(j\omega)_{\min}$.

* Decimal classification: R143. Original manuscript received by the IRE, January 19, 1954.

† Div. of Elec. Eng., Univ. of Calif., Berkeley 4, Calif.

[1] Since, when including the point at infinity, every rational function has as many poles as zeros, we need only refer to the number of poles.

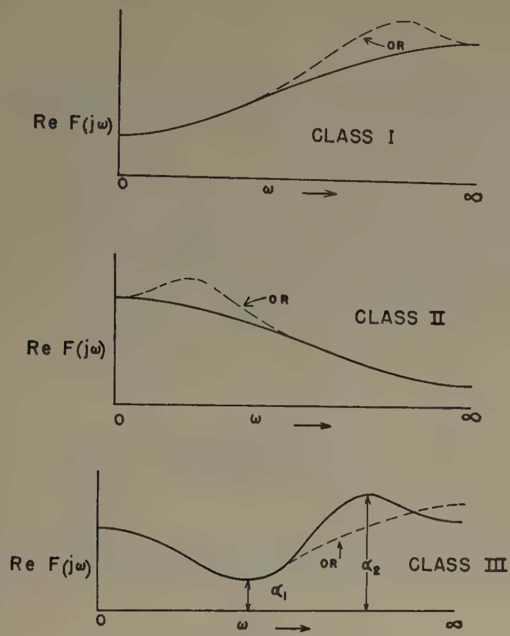
[2] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y.; 1945.

[3] O. Brune, "Synthesis of a finite two-terminal network whose driving-point impedance is a prescribed function of frequency," *Jour. Math. & Phys.*, vol. 10, pp. 191–236; 1931.

[4] Herein we shall abbreviate *positive-real* as *p-r*.

[5] E. A. Guillemin, "Modern Methods of Network Synthesis," *Advances in Electronics*, Edited by L. Marton, Academic Press, Inc., New York, N. Y., vol. III, pp. 261–303; 1951.

at $\omega=0$; Class II having $Re\ F(j\omega)_{\text{min.}}$ at $\omega=\infty$; and Class III having $Re\ F(j\omega)_{\text{min.}}$ at finite values of $\pm\omega$ (see Figs. 1, 2, and 3, respectively).



Figs. 1, 2, and 3—Real part characteristics which depict three classes of positive-real functions.

An important process in the synthesis of networks is the procedure of removing a constant from the immittance function so that the resulting function $F'(p)$ has the property that

$$Re\ F'(j\omega)_{\text{min.}} = 0. \tag{6}$$

Immittance functions which satisfy (6) are often called minimum-resistance, minimum-conductance functions; but here we shall refer to them as *minimum-real-part functions*.

It is often necessary to determine how large a constant must be subtracted from a function to make the remainder satisfy (6). If all of the coefficients in (1) are nonzero, then if $F(p)$ is a Class I immittance

$$Re\ F(j\omega)_{\text{min.}} = F(j0) = b/d. \tag{7}$$

Thus for this case,

$$F'(p) = F(p) - b/d \tag{8}$$

will be a minimum real-part, p-r function.

Similarly if $F(p)$ is Class II and all of the coefficients in (1) are nonzero,

$$Re\ F(j\omega)_{\text{min.}} = F(j\omega) = g, \tag{9}$$

and removal of a constant equal to g will give the desired result.

For the Class III case we may utilize the following analysis. Using (3), let

$$Re\ F'(j\omega) = Re\ F(j\omega) - \alpha_k$$

$$= \left.\frac{m_1m_2 - n_1n_2}{m_2^2 - n_2^2}\right|_{p=j\omega} - \alpha_k$$

$$= \left.\frac{m_1m_2 - n_1n_2 - \alpha_k m_2^2 + \alpha_k n_2^2}{m_2^2 - n_2^2}\right|_{p=j\omega}. \tag{10}$$

When $F'(p)$ is a Class III, minimum real-part function; from (10) it can be seen that

$$m_1m_2 - n_1n_2 - \alpha_k m_2^2 + \alpha_k n_2^2 = 0 \tag{11}$$

(a polynomial in p) must have *second-order* roots on the $j\omega$ axis at the frequencies where (6) is satisfied. By analysis of (11) it can be shown that double roots will appear on the $j\omega$-axis for values of α_k which satisfy the equations:

$$(U^2-4d^2)\alpha_k^2+(4gd^2+4bd-2UW)\alpha_k+(W^2-4gbd)=0 \tag{12}$$

and

$$x_k = \frac{W - \alpha_k U}{2g - 2\alpha_k} > 0, \tag{13}$$

where

$$W = b - ae + dg,$$

and

$$U = 2d - e^2,$$

the a, b, e, d, and g parameters being those of (1). It is possible for there to be two values of α_k which will satisfy (12) and (13). Then one of the values is the height of the $Re\ F(j\omega)$ curve at a minimum while the other is the height at a maximum as illustrated by α_1 and α_2 in Fig. 3. Thus the function

$$F'(p) = F(p) - \alpha_1 \tag{14}$$

will be a Class III, minimum real-part p-r function.

Sometimes it is necessary to know at what frequency the minimum or maximum of $Re\ F(j\omega)$ occurs for Class III functions. These frequencies are easily computed from

$$j\omega_k = j\sqrt{x_k}, \tag{15}$$

where x_k is defined by (13).

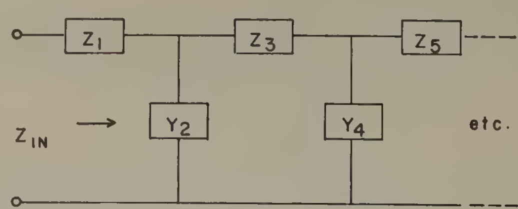


Fig. 4—Impedance realization of a continued fraction of positive-real terms.

Continued-Fraction Expansions

The input impedance function of the ladder network of Fig. 4 may be represented by the continued-fraction expansion

$$Z_{\text{in}} = Z_1 + \cfrac{1}{Y_2 + \cfrac{1}{Z_3 + \cfrac{1}{Y_4 + \cdots \text{ etc.}}}} \cdot \qquad (16)$$

Also the input admittance of the network in Fig. 5 may analogously be represented by

$$Y_{\text{in}} = Y_1 + \cfrac{1}{Z_2 + \cfrac{1}{Y_3 + \cfrac{1}{Z_4 + \cdots \text{ etc.}}}} \cdot \qquad (17)$$

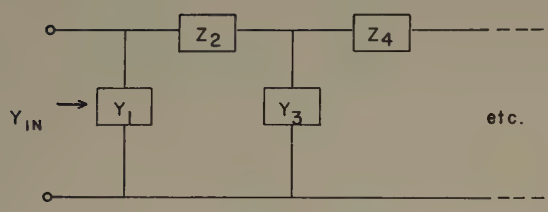


Fig. 5—Admittance realization of a continued fraction of positive-real terms.

Some kinds of immittances are very easily realized in these forms by making a continued-fraction expansion by use of what shall herein be referred to as *forward* and *reverse division.* Let

$$F_0(p) = \frac{g_n p^n + g_{n-1}p^{n-1} + \cdots + g_1 p + g_0}{h_k p^k + h_{k-1}p^{k-1} + \cdots + h_1 p + h_0} \qquad (18)$$

be an immittance of arbitrary complexity.[6] We shall define *forward division* as long division carried out in the manner indicated by:

$$h_k p^k + \cdots + h_1 p + h_0 \overline{) g_n p^n + \cdots + g_1 p + g_0}. \qquad (19)$$

Similarly *reverse division* will be defined as long division carried out in the manner indicated by:

$$h_0 + h_1 p + \cdots + h_k p^k \overline{) g_0 + g_1 p + \cdots + g_n p^n}. \qquad (20)$$

An important kind of continued-fraction expansion can be made by use of the following operations involving forward and reverse division:

1. If $g_0 > 0$ but $h_0 = 0$, then $F_0(p)$ has a pole at the origin, and it can be removed by one step of reverse division. The result will be of the form

$$F_0(p) = \frac{g_0}{h_1 p} + F_1(p), \qquad (21)$$

where $F_1(p)$ will not have a pole at the origin and will be *p-r* if $F_0(p)$ is *p-r*.

2. If $g_0 = 0$, but $h_0 > 0$, then $F_0(p)$ has a zero at the origin. In this case the function is inverted to make the zero at the origin become a pole and then operation 1 is applied to remove the pole. In this case the result is of the form

$$F_0(p) = \cfrac{1}{\cfrac{h_0}{g_1 p} + F_2(p)}, \qquad (22)$$

where $F_2(p)$ will not have a pole at the origin and will be *p-r* of $F_0(p)$ is *p-r*.

3. If terms g_0 and h_0 are present and $F_0(p)$ is a Class I *p-r* function, then a step of reverse division will remove a constant equal to $Re\ F(j\omega)_{\text{min.}} = Re\ F(j0) = F(0)$, and the remainder function $F_3(p)$ will be a minimum real-part *p-r* function. The result is of the form

$$F_0(p) = \frac{g_0}{h_0} + F_3(p). \qquad (23)$$

The remainder function $F_3(p)$ will necessarily have a zero at the origin so it can be broken down further by operation 2.

4. If $n = k+1$, then $F_0(p)$ has a pole at infinity which can be removed by forward division to give

$$F_0(p) = \frac{g_n p}{h_k} + F_4(p). \qquad (24)$$

$F_4(p)$ will not have a pole at infinity and will be *p-r* if $F_0(p)$ is *p-r*.

5. If $n = k-1$, then $F_0(p)$ has a zero at infinity. In this case the function is inverted to make the zero a pole, and then operation 4 is applied to give

$$F_0(p) = \cfrac{1}{\cfrac{h_k p}{g_n} + F_5(p)} \cdot \qquad (25)$$

$F_5(p)$ will not have a pole at infinity and will be *p-r* if $F_0(p)$ is *p-r*.

6. If $F_0(p)$ is Class II function and $n = k$, then a step of forward division will remove a constant equal to $Re\ F_0(j\omega)_{\text{min.}} = Re\ F(j\infty) = F(\infty)$, and the remainder function $F_6(p)$ will be a minimum real-part function. The result is of the form

$$F_0(p) = \frac{g_n}{h_k} + F_6(p). \qquad (26)$$

The function $F_6(p)$ will necessarily have a zero at infinity so it in turn can be broken up by use of operation 5.

By use of these six operations, numerous immittance functions can be completely broken into a continued fraction of simple *p-r* terms. These terms can then be identified as series impedances or shunt admittances of a ladder network in accordance with (16) or (17). Many readers will recognize that Cauer's continued-fraction method for synthesis of RC, RL, and LC immittances

[6] Since $F_0(p)$ in (18) is to be *p-r*, the highest powers in the numerator and denominator polynomials cannot differ by more than one and likewise for the lowest powers.

utilizes these operations.[7] These six operations are, however, also useful for synthesis of some RLC networks. The necessary condition which makes it possible to completely expand an immittance this way is that after each step of forward or reverse division, the remainder function or its reciprocal must be a Class I or II *p-r* function. If the remainder function and its reciprocal are both Class III, additional techniques must be introduced in order to break the remainder function down.

Synthesis of Minimum Real-Part, Class I and II Imittances Having Two Poles

To illustrate the use of the forward and reverse division operations, let us consider the synthesis of Class I and II minimum real-part functions having two poles. Any two-pole, minimum real-part, Class I function can be written in the form

$$F(p) = \frac{gp^2 + ap}{p^2 + ep + d} \tag{27}$$

by letting $b=0$ in (1). If *p-r* condition A is satisfied, we find from (5) that *p-r* condition B will be satisfied for this case if

$$a/g \geqq d/e. \tag{28}$$

Thus it is easy to check (27) for *p-r* character. The function

$$F(p) = \frac{p^2 + 3p}{p^2 + 2p + 2} \tag{29}$$

is of the form of (27) and satisfies (28). Since (29) has a zero at the origin, we can break $F(p)$ up some by applying operation 2 to give

$$F(p) = \frac{1}{\dfrac{2}{3p} + \dfrac{4/3 + p}{3 + p}}. \tag{30}$$

The remainder function,

$$F_2(p) = \frac{4/3 + p}{3 + p}, \tag{31}$$

has only one pole and one zero. All *p-r* functions with only one pole and one zero must be either Class I or Class II functions. Evaluation of (31) at zero and infinity shows that $F_2(p)$ is a Class I function. By operation 3 the expansion becomes

$$F(p) = \frac{1}{\dfrac{2}{3p} + 4/9 + \dfrac{(5/9)p}{3 + p}}. \tag{32}$$

Inverting the last term and dividing again gives

$$F(p) = \frac{1}{\left(\dfrac{2}{3p} + 4/9\right) + \dfrac{1}{\left(\dfrac{27}{5p} + \dfrac{9}{5}\right)}}. \tag{33}$$

It is interesting to note that a somewhat different expansion can be obtained by inverting the $F_2(p)$ part of (30) before breaking it up. Inverting $F_2(p)$ and using operation 6 leads to the expansion:

$$F(p) = \frac{1}{\dfrac{2}{3p} + \dfrac{1}{1 + \dfrac{1}{\dfrac{3}{5}p + \dfrac{1}{5/4}}}}. \tag{34}$$

If $F(p)$ is an impedance, then expansions (33) and (34) can be identified with (16) and Fig. 4. Equation (33) gives the circuit shown in Fig. 6(a) while (34) gives the circuit in Fig. 6(b). Note that for these particular expansions the Z_1 impedance is zero. If $F(p)$ is defined as an admittance, expansions (33) and (34) may be identified with (17) and Fig. 5, and the "inverse" networks to those in Figs. 6(a) and 6(b) will be obtained.

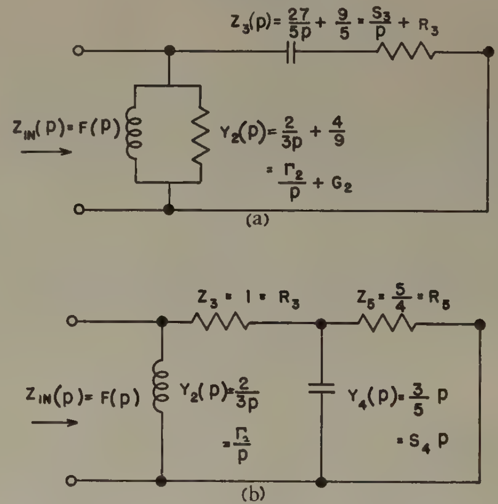


Fig. 6—(a) An impedance realization of a Class I, minimum real part function having two poles (a special case classifiable as either a Type *A* or *B* realization). (b) A realization equivalent to that in (a).

Any Class II, minimum real-part function with two poles can be expressed in the form

$$F(p) = \frac{ap + b}{p^2 + ep + d}. \tag{35}$$

This is simply (1) with $g=0$. Equation (5) shows that if *p-r* condition A is fulfilled, then function (35) is *p-r* if

$$b/a \leqq e. \tag{36}$$

Function (35) is also easily expanded by use of forward and reverse division, and the network can again be found by identifying the expansion with (16) or (17) and Figs. 4 or 5, respectively. In general no more than

[7] E. A. Guillemin, "Communication Networks," John Wiley & Sons, Inc., New York, N. Y., vol. II, pp. 198–216; 1935.

four elements will be required to realize a Class I or II, minimum real-part function with two poles.

The General Case with Two Poles and Zeros

Now we shall review procedures which taken together provide means for realizing any immittance having the form of (1).

Type A Realization: This procedure works for Class I or II immittances only. If in (1)

$$g > b/d, \tag{37}$$

then $F(p)$ *may* possibly be a Class I function. As is shown in Appendix A, $F(p)$ is a Class I, *p-r* function if (37) and *p-r* condition A are satisfied while

$$\frac{ad - be}{gd - b} \geqq \frac{d}{e} \cdot \tag{38}$$

If in (1)

$$g < b/d, \tag{39}$$

then $F(p)$ is possibly a Class II function. Appendix A shows that $F(p)$ is a Class II, *p-r* function if (39) and *p-r* condition *A* are satisfied while

$$0 < \frac{b - gd}{a - ge} \leqq e. \tag{40}$$

If $F(p)$ is a Class I function (nonminimum real-part) then a realization is easily obtained by making a continued-fraction expansion starting with reverse division (operation 3) and then continuing with the forward or reverse division operations that become appropriate. The same is true for the Class II case, only then the expansion begins with forward division (operation 6).

For a numerical example consider the function

$$F(p) = \frac{22p^2 + 46p + 45}{44p^2 + 48p + 54} \cdot \tag{41}$$

This function is seen to satisfy (39) and (40), hence it is a Class II, nonminimum real-part function. One possible form for the continued-fraction expansion is

$$F(p) = \frac{1}{2} + \cfrac{1}{2p + \cfrac{1}{\cfrac{1}{3} + \cfrac{1}{\cfrac{3}{p} + \cfrac{1}{(3/2)}}}} \cdot \tag{42}$$

Supposing $F(p)$ to be an admittance, by identifying (42) with (17) and Fig. 5 we get the network shown in Fig. 7.

$F(p)$ functions of the form of (1) which satisfy *p-r* condition A and (37) and (38) or (39) and (40) are all susceptible to this type of realization which will be herein referred to as a Type A realization. Type A realizations will require no more than five elements, though some special cases will require less.

Type B Realization: Some $F(p)$ functions which are Class III as they stand become Class I or II functions if inverted. If $F(p)$ fails to qualify for a Type A realization we may next test for the properties

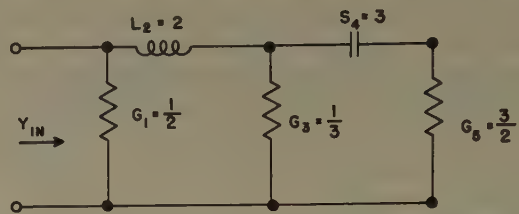


Fig. 7—Admittance realization of a nonminimum real part, Class II function having two poles (Type *A* realization).

$$g > b/d \quad \text{and} \quad 0 < \frac{b - dg}{a - eg} \leqq a/g \tag{43}$$

or

$$g < b/d \quad \text{and} \quad \frac{eb - da}{b - dg} \geqq b/a, \tag{44}$$

(*p-r* condition A assumed to be satisfied). If conditions (43) are satisfied the reciprocal of $F(p)$ is a Class II function while if conditions (44) are satisfied the reciprocal of $F(p)$ is a Class I function. To get the network we may start with

$$F(p) = \frac{1}{1/F(p)}, \tag{45}$$

and then expand $1/F(p)$ as in a Type A realization. Thus (45) will lead to a continued fraction which is without what would ordinarily be the first term. This was also the case in (33) and (34), and causes no trouble. Type B realizations also require no more than five elements.

It should be noted that sometimes when $F(p)$ is Class I or II, $1/F(p)$ may be respectively Class II or I, hence both Type A and B realizations may be possible.

Type C Realization: If neither $F(p)$ nor its reciprocal is a Class I or II function then the Type A and B procedures fail. Any Class III function can be realized by Brune's method,[8] but his method requires unity-coupled coils in order to realize a Class III function of the form of (1).[8] If unity-coupled coils are to be excluded the simplest procedure appears to result from breaking the Class III function into the sum of a Class I, minimum real-part function plus a Class II, minimum real-part function. This gives

$$F(p) = \frac{gp^2 + tp}{p^2 + ep + d} + \frac{(a - t)p + b}{p^2 + ep + d} \cdot \tag{46}$$

It can be shown that both terms in (46) will be *p-r* if *p-r* condition A is satisfied and

$$gd/e \leqq t \leqq a - b/e. \tag{47}$$

[8] Brune also shows how to obtain what were herein called Type A and B realizations. The Type A and B procedures described in this paper differ from his in mathematical technique only.

Consider the example:

$$F(p) = \frac{p^2 + 4p + 7}{p^2 + 2p + 1}. \tag{48}$$

In this case (47) can be satisfied on both sides by an equal sign if $t = 1/2$. Equation (46) becomes

$$F(p) = \frac{p^2 + 0.5p}{p^2 + 2p + 1} + \frac{3.5p + 7}{p^2 + 2p + 1}. \tag{49}$$

The two terms in (49) can be expanded in continued fractions to give

$$F(p) = \cfrac{1}{\cfrac{2}{p} + \cfrac{1}{1 + \cfrac{1}{2p}}} + \cfrac{1}{\cfrac{1}{3.5}p + \cfrac{1}{3.5p + \cfrac{1}{(1/7)}}}. \tag{50}$$

Equation (50) represents two, ladder networks connected together. If $F(p)$ is construed to be an impedance, the realization is as shown in Fig. 8.

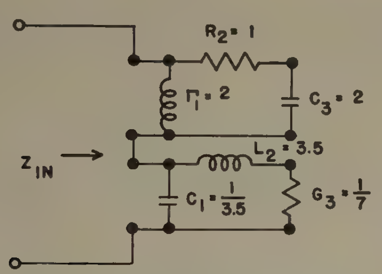


Fig. 8—A Type *C* realization example.

Observe that Fig. 8 represents two, three-element ladders connected in series. In most cases it will only be possible to satisfy (47) by an equal sign on one side. Then one of the ladders will have four elements while the other will have only three. In general it will be found that this type of realization will require no more than seven elements. It can be shown that if $F(p)$ fails to qualify for Type C realization, then $1/F(p)$ also fails to qualify.

Type D Realization: If $F(p)$ satisfies *p-r* condition A and (5), but doesn't satisfy the conditions for Type A, B, or C realization, then what will herein be called a Type D realization may be utilized to obtain a circuit without unity-coupled coils. In these cases Type C realization fails because $Re\, F(j\omega)_{\min.}$ is so small that the Class III, $F(p)$ function cannot be broken into the sum of a *p-r* Class I function plus a *p-r* Class II function as specified by (46) and (47).

The first step of this procedure is to break $F(p)$ up into a constant plus a minimum real-part function. Thus by (10) to (14) we can obtain

$$F(p) = \alpha_1 + F'(p), \tag{51}$$

where $F'(p)$ is a function such that $Re\, F'(j\omega) = 0$ at $j\omega = \pm j\omega_1$. The next step is to break $F'(p)$ up into a term having a zero at infinity and zeros at $p = \pm j\omega_1$; and a second term having a zero at the origin and zero real-part at $p = \pm j\omega_1$. This may be accomplished by multiplying the numerator and denominator of $F'(p)$ by $(p+K)$ to give[9]

$$F'(p) = \frac{(gp^2 + ap + b)(p + K)}{(p^2 + ep + d)(p + K)}, \tag{52}$$

and then splitting (52) to yield

$$F'(p) = \frac{J(p + j\omega_1)(p - j\omega_1)}{(p^2 + ep + d)(p + K)} + \frac{(gp^2 + ap + b)(p + K) - J(p + j\omega_1)(p - j\omega_1)}{(p^2 + ep + d)(p + K)}. \tag{53}$$

It can be shown that the first term will be *p-r* if

$$K = \frac{e\omega_1^2}{d - \omega_1^2} \tag{54}$$

while the second term will be *p-r* and will have a zero at the origin if K has the above value and

$$J = \frac{bK}{\omega_1^2}. \tag{55}$$

The final result is:

$$F'(p) = \frac{Jp^2 + J\omega_1^2}{p^3 + (e + K)p^2 + (d + Ke)p + dK} + \frac{gp^3 + (a + gK - J)p^2 + (b + aK)p}{p^3 + (e + K)p^2 + (d + Ke)p + dK}. \tag{56}$$

The first term of (56) can be expanded in a continued fraction by starting with inversion and forward division (operation 5) while the second term can be expanded by starting with inversion and reverse division (operation 2).

Consider the example

$$F(p) = \frac{1.1p^2 + 0.8130p + 0.29394}{p^2 + 2p + 2}. \tag{57}$$

Both function (57) and its reciprocal are Class III functions. The coefficients in (57) fail to satisfy (47) so Type D realization is necessary.

We may use (12), (13), and (15) to find $\alpha_1 = 0.1$ and $j\omega_1 = j0.6588$. By (14)

$$F'(p) = \frac{p^2 + 0.6130 + 0.09394}{p^2 + 2p + 2}. \tag{58}$$

By (54) and (55), $K = 0.5542$ and $J = 0.1197$, respectively. Equation (56) gives

$$F'(p) = \frac{0.1197p^2 + 0.05206}{p^3 + 2.554p^2 + 3.109p + 1.109}$$

[9] For the remainder of this Type D realization discussion the g, a, b, e, and d parameters will be used to refer to $F'(p)$.

$$+\frac{p^3+1.047p^2+0.4337p}{p^3+2.554p^2+3.109p+1.109}. \quad (59)$$

Adding α_1 to (59) and expanding in two, continued fractions we obtain

$$F(p)=0.1+\cfrac{1}{(8.355p+21.34)+\cfrac{1}{\left(0.04477p+\cfrac{0.01945}{p}\right)}}$$
$$+\cfrac{1}{\cfrac{2.5560}{p}+\cfrac{1}{1+\cfrac{1}{0.9548p+\cfrac{1}{2.423p}}}}. \quad (60)$$

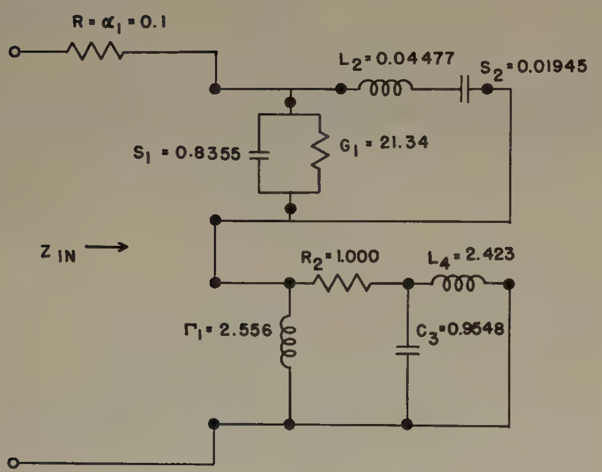


Fig. 9—A Type *D* realization example.

If $F(p)$ is taken to be an impedance then (60) represents the circuit shown in Fig. 9. This circuit has nine elements which is the maximum number required for a Type D realization. Some readers will note that this same circuit could be obtained by the method of Bott and Duffin.[10] However, this approach is believed to be somewhat easier since the Bott and Duffin method requires finding the real, positive root of a third-order polynomial.[11]

The Ranges of Realization

To give a better insight into when these different kinds of realization are possible, Fig. 10 illustrates the various ranges of zero locations of $F(p)$, (1), when the poles are located at $p=-2\pm j1$. Only the second quadrant of the p-plane is shown since no zeros can occur in the first or fourth quadrants and the second and third quadrants are symmetrical with respect to the real axis. With these given poles, $F(p)$ will be *p-r* if it has a real, positive multiplier and the zeros occur in conjugate pairs such that the second-quadrant zero lies within the outer curved contour. If the zero lies within the cross-hatched region marked A_I (45 degree cross-hatch lines falling from left to right), then the function is Class I and can be realized with a Type A realization. If the zero is in the A_{II} region (45 degree cross-hatch rising from left to right), $F(p)$ is a Class II function and Type A realization is again possible. Note that the region of Type B realization (horizontal cross-hatching) overlaps the Type A region in places. Where they overlap, both $F(p)$ and its reciprocal are Class I or II functions; and where they do not overlap, either $F(p)$ or its reciprocal is a Class III function. If the second-quadrant zero lies within the outer contour but outside of the regions of Type A or B realization, then both $F(p)$ and its reciprocal are Class III *p-r* functions.

The region of Type C realizability overlaps all of the region of Type A realization and part of the region of Type B realization; however, since more elements are required, one would probably only want to use Type C realization when the zero lies in one of the unshaded regions marked *C*. The region of Type D realization overlaps part of the region of Type C realization and part of the region of Type B; but again since Type D realization requires the most elements of all, it would usually be undesirable unless the second-quadrant zero lies in one of the vertically cross-hatched regions marked *D*. If $F(p)$ has simple zeros on the real axis, any of the Type A, B, C, or D realizations may be possible depending on relative locations of zeros. It is interesting that the boundaries of various regions of realization in Fig. 10 all have geometric symmetry with respect to the circle about the origin which passes through the poles.

Conclusion

Immittances having two poles are commonly used as building blocks in the design of filters, equalizers, and amplifiers. By use of the formulas presented in this paper it is possible to relatively quickly determine the significant properties of such immittances and choose an appropriate synthesis procedure. Continued-fraction expansions based on forward and reverse division are seen to provide a straight-forward, yet flexible tool for synthesis of RLC networks, as well as networks with only two kinds of elements. The Type A, B, C, and D synthesis procedures described are also suggestive of the synthesis methods which are possible when the immittance has more than two poles and zeros.

Appendix A

If (1) is a Class I function then

$$Re\,F(j\infty)=g>Re\,F(j0)=Re\,F(j\omega)_{\min.}, \text{ so}$$
$$F'(p)=F(p)-b/d$$
$$=\frac{\left(g-\frac{b}{d}\right)p^2+\left(a-\frac{be}{d}\right)p}{p^2+ep+d}$$

[10] R. Bott and R. J. Duffin, "Impedance synthesis without the use of transformers," *Jour. Appl. Phys.*, vol. 20, p. 816; August, 1949. Also "Modern methods of network synthesis," ibid., part XVI.
[11] "Modern methods of network synthesis," ibid., p. 294.

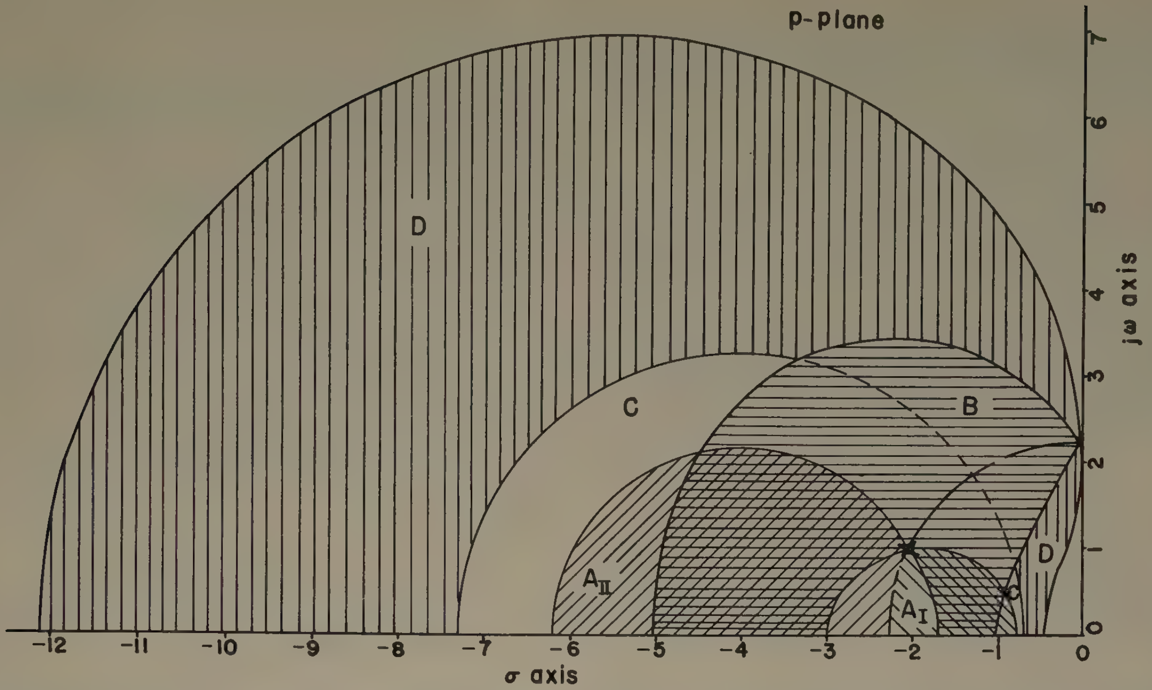


Fig. 10—Ranges of second-quadrant zero locations for the various kinds of realizations when the immittance-function poles are at $p=-2\pm j1$.

$$= \frac{g'p^2 + b'p}{p^2 + ep + d} \tag{61}$$

must be a Class I, minimum real-part, p-r function. By applying condition (28) to (61), condition (38) is obtained.

If (1) is a Class II function then

$$Re\,F(j\omega)_{\min.} = Re\,F(j\infty) = g < Re\,F\,(j0) = b/d, \quad \text{so}$$

$$F''(p) = F(p) - g$$

$$= \frac{(a - ge)p + (b - dg)}{p^2 + ep + d} = \frac{a''p + b''}{p^2 + ep + d}. \tag{62}$$

must be a Class II, minimum real-part, p-r function. Application of condition (36) to (62) yields condition (40).

Part II: A Constant-Resistance Ladder for Transfer Function Synthesis

Generation of Points of Infinite Gain and Loss

If a transfer function is defined so that

$$T(p) = \frac{\text{Input}}{\text{Output}}, \tag{63}$$

then the zeros of $T(p)$ will be points of infinite gain and will occur at frequencies of natural vibration of the circuit, while the poles will be points of infinite loss. If the system is driven by a zero internal impedance voltage generator, then the frequencies of natural vibration will occur at the same frequencies as the zeros of the input impedance. Except for possible cancellation effects, the zeros of the input impedance will be the same as the zeros of $T(p)$.[12] On the other hand if the system is driven by an infinite internal impedance current generator, then the frequencies of natural vibration will correspond with the poles of the input impedance. In this case the zeros of $T(p)$ will be the same as the poles of the input impedance except for possible exceptions due to cancellation between poles and zeros in the input impedance or in the transfer function. Thus we see that in many cases the zeros of $T(p)$ will be completely defined when the input impedance of the system is defined.

The poles of $T(p)$ are points at which all transmission is stopped. They arise physically in the following ways:

1. If part of the circuit has more than one route of transmission, points of infinite loss (poles of $T(p)$) will be generated at frequencies at which the ouputs from the various routes cancel.
2. By the poles of series impedances (such as Z_1 in the ladder of part I, Fig. 4) blocking transmission, provided that the impedance of the network to the right of Z_1 does not have the same poles. When Z_1 and the impedance to the right have the same poles they act as a voltage divider and transmission is not blocked.
3. By the poles of shunt admittances (such as Y_1 in the ladder of part I, Fig. 5) shorting out the signal, provided that the admittance of the network to the right of Y_1 does not have the same poles. If Y_1 and the network to the right have the same poles of admittance then they will act as a current divider and transmission will not be blocked.

[12] Sometimes the presence of some input-impedance zeros is obscured because they are cancelled by poles at the same frequencies. These cancelled zeros are nevertheless indicative of natural modes which will be apparent in other meshes, and they will correspond to zeros of $T(p)$ if there are no further cancellations between transfer function poles and zeros.

The points of infinite loss of lattice, bridged-T, and parallel-ladder networks are generated primarily by the first method. Having only one route of transmission, a single-ladder network generates its points of infinite loss entirely by the second and third methods.

Synthesis of a Constant-Resistance Ladder

Let us now consider the design of a ladder network having a constant-resistance input and a prescribed, minimum-phase transfer function. We shall stipulate that the network is to be driven by a generator with a one-ohm internal impedance, and ladder impedance is to match this. Transfer function (63) may be written

$$T(p) = A_1 \frac{H_1}{H_2} = A_1 T_m(p) \tag{64}$$

where H_1 and H_2 are polynomials, A_1 is a constant, and $T_m(p)$ will be defined later. We shall temporarily stipulate that $T(p)$ is a nonminimum real part p-r function; however as we shall see, this stipulation is easily removed.

If the generator has an internal resistance of one ohm and the network has an input resistance of one ohm, then the zero-impedance voltage generator in a Thevenin equivalent circuit will see the resistance:

$$Z_{\text{in}} = 2 = 2\frac{H_1}{H_1} = 1 + \frac{H_1}{H_1}. \tag{65}$$

The polynomial H_1 has been introduced in (65) to give the circuit the proper frequencies of natural vibration. Now let us define

$$\nu_1 = Re \left.\frac{H_1(j\omega)}{H_2(j\omega)}\right|_{\text{min.}} > 0 \tag{66}$$

and

$$\nu_2 = Re \left.\frac{H_2(j\omega)}{H_1(j\omega)}\right|_{\text{max.}} > 0. \tag{67}$$

Both ν_1 and ν_2 will be greater than zero since $T(p)$ was stipulated to be a nonminimum real-part p-r function. Then (65) can be written

$$\begin{aligned} Z_{\text{in}} &= 1 + \left(\frac{H_1}{H_1} - \frac{H_2}{K_2H_1}\right) + \frac{H_2}{K_2H_1} \\ &= 1 + Z_1 + Z_a, \end{aligned} \tag{68}$$

where

$$K_2 \geqq \nu_2. \tag{69}$$

Both Z_1 and Z_a are guaranteed to be p-r due to (67), (69), and the stipulation that (64) is p-r.

Equation (68) can also be expressed as the continued-fraction:

$$\begin{aligned} Z_{\text{in}} &= 1 + \left(\frac{K_2H_1 - H_2}{K_2H_1}\right) + \cfrac{1}{\cfrac{K_2H_1}{H_2}} \\ &= 1 + Z_1 + \frac{1}{Y_a}. \end{aligned} \tag{70}$$

Since (64) was stipulated to be a nonminimum real-part p-r function, a shunt conductance can be removed from Y_a in (70) to give:

$$\begin{aligned} Z_{\text{in}} &= 1 + \left(\frac{K_2H_1 - H_2}{K_2H_1}\right) + \cfrac{1}{\left(\cfrac{K_2H_1}{H_2} - K_1K_2\right) + K_1K_2} \\ &= 1 + Z_1 + \frac{1}{Y_2 + G_L} \end{aligned} \tag{71}$$

where

$$0 < K_1 \leqq \nu_1. \tag{72}$$

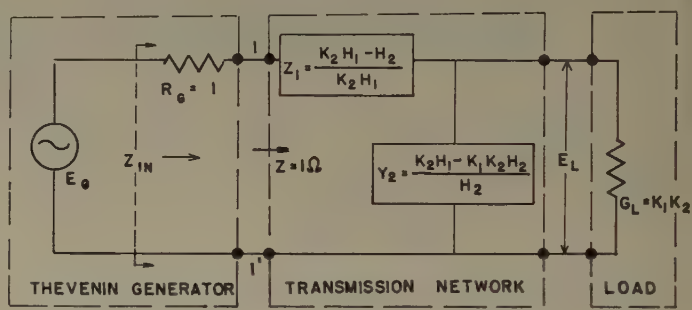


Fig. 11—A constant-resistance ladder section.

Equation (71) corresponds to the circuit in Fig. 11. Note that (65) causes this network to have natural frequencies of vibration at the zeros of H_1 when Z_{in} is driven by *either* a zero-impedance voltage generator or an infinite-impedance current generator. Since the zeros of $T(p)$ are natural frequencies of vibration, the transfer function for this network has the same zeros as those in (64). By (68), since the impedance Z_a seen to the right of Z_1 has the same poles as Z_1, the impedance Z_1 serves as what Guillemin and others refer to as a "zero-shifting branch," and it does not create any points of infinite loss.[13] However, points of infinite loss will be created at the poles of Y_2 and these are seen to be the same as the poles (points of infinite loss) of the transfer function (64). The network in Fig. 11 has the transfer function (64) within a constant multiplier. Carrying out the synthesis in a similar but dual manner, constant-resistance network in Fig. 12 is obtained. This network also has the transfer function (64) within a constant multiplier.

Since the input impedance of the transmission network in Fig. 11 (or in Fig. 12) is a constant-resistance, the impedance level of one section of this type can be adjusted so that its input will serve as the proper constant load resistance for another section. In this manner any number of such simple sections can be designed separately and then cascaded together to give a ladder

[13] "Modern methods of network synthesis," ibid., pp. 286–290.

network whose transfer function is the product of the section transfer function.[14] Thus as with other constant-resistance networks,[15] the realization of a complicated transfer function can be greatly simplified by realizing it in a number of simple parts. It should be noted that the transfer function for each component section must be a nonminimum real-part p-r function, *but the over-all transfer function need not be p-r!* The process then is to select poles and zeros for each section of the ladder so that the section transfer function will be nonminimum real-part p-r. If it should be impossible to factor the over-all transfer function into a complete set of p-r factors, additional factors of the form $(p-p_1)/(p-p_1)$ can always be introduced so as to make p-r factorization possible. Thus any minimum phase transfer function can be realized within a flat-loss factor in a constant-resistance ladder.

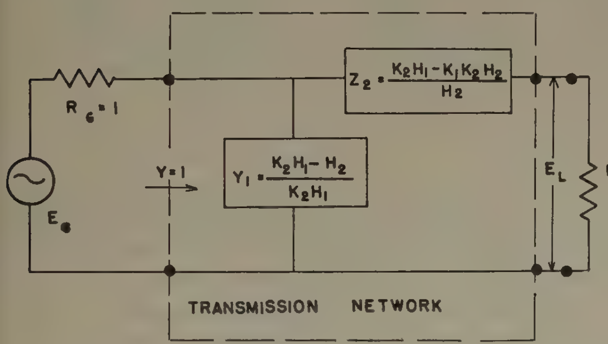


Fig. 12—An alternate constant-resistance ladder section.

If the over-all transfer function is factored so that each section transfer function has either one or two poles and zeros, then the techniques discussed in Part I can be used to simplify the calculations considerably. Equations (38), (40), and (5) can be used to determine if the section transfer functions are Class I, II, or III p-r functions; and then (7), (9), (12), and (13) can be adapted for computation of ν_1 and ν_2. The branch immittances will have no more than two poles and zeros and therefore can be relatively quickly synthesized by the methods of Part I.

Constant-Resistance Ladder Flat-Loss Comparison

Since these networks will often have unequal terminations, let us define our transfer function in terms of the voltage available at the load conductance G_L (or resistance R_L). Thus,

$$P_{g\ \text{Avail.}} = \frac{E_g^2}{4R_g} = P_{L\ \text{Avail.}} = G_L E_L^2{}_{\text{Avail.}}, \quad (73)$$

where $P_{g\ \text{Avail.}}$ is the available power of the generator, E_g and R_g are as defined in Fig. 11, $P_{L\ \text{Avail.}}$ is the power available at the load, and $E_{L\ \text{Avail.}}$ is the voltage which this generator would cause across the given load G_L if the generator and load were perfectly matched by an ideal transformer. From (73),

$$E_{L\ \text{Avail.}} = \frac{1}{2\sqrt{R_g G_L}} E_g. \quad (74)$$

It appears logical to define the transfer function then as

$$T(p) = \frac{E_{L\ \text{Avail.}}}{E_L}, \quad (75)$$

the ratio of the available load voltage to the delivered load voltage.[16] Since the magnitude of the delivered load voltage cannot exceed that of the available voltage across any given G_L,

$$|T(j\omega)| = \left|\frac{E_{L\ \text{Avail.}}}{E_L}\right| \geqq 1. \quad (76)$$

Any transfer function (75) that satisfies (76) with an equal sign at any steady-state frequency $j\omega$ is a minimum-loss transfer function. Let us define

$$T_m(p) = \frac{1}{H_2} \quad (77)$$

as a minimum-loss transfer function. Then A_1 in (64) is a constant factor, equal to or greater than one, which indicates the flat loss of the network. For the networks of Figs. 11 or 12, A_1 is computed to be

$$A_1 = \sqrt{\frac{K_2}{K_1}}. \quad (78)$$

The smallest value for A_1 (and the least flat loss) will be obtained when (69) and (72) are satisfied with equal signs to give:

$$A_1|_{\text{min.}} = \sqrt{\frac{\nu_2}{\nu_1}}. \quad (79)$$

Since any transfer function which can be realized in one conventional, RLC, constant-resistance bridged-T section[17] can also be realized in one constant-resistance ladder section, the two have about the same realm of application, and it appears relevant to compare their attenuation factors. The smallest attenuation factor for the conventional, RLC bridged-T is computed to be:

$$A_2|_{\text{min.}} = \frac{1}{\nu_1}. \quad (80)$$

Since transfer function (77) has no flat loss, from (66) and (67) it can be seen that both ν_1 and ν_2 must be no greater than one. Hence the ladder $A_{1\ \text{min.}}$ will always be less than the bridged-T $A_{2\ \text{min.}}$ except when both methods of realization give minimum-loss transfer function, i.e., except when $A_1|_{\text{min.}} = A_2|_{\text{min.}} = 1$.

[14] Depending on how the transfer function is defined, the over-all transfer function may equal the product of the parts, only within a constant multiplier.
[15] Bode, op. cit., chap. 12.

[16] Equation (75) is in nature an input/output ratio and is consistent with the definition of $T(p)$ in (63).
[17] Bode, op. cit., pp. 270–275.

Some Practical Considerations

Let us assume that the poles and zeros of a complicated transfer function are to be factored into groups and then realized in a chain of constant-resistance ladder sections. Let us suppose further that each section is to contribute two poles and zeros to the over-all transfer function. It may be possible to group the poles and zeros so that most or all of the section transfer functions are Class I or II *p-r* functions. Each section that has a Class I or II transfer function will have Class I or II immittances in *both* arms, and considerably fewer elements will be required than if Class III immittances were involved. Let us suppose

$$\frac{gp^2 + ap + b}{p^2 + ep + d} = \frac{\text{Input}}{\text{Output}} \tag{81}$$

has been factored out to be realized as one section of a ladder. It can be shown that if (81) satisfies (37) and (38) of Part I, then (81) and all of the arm immittances in Figs. 11 and 12 will be Class I functions. Then the corresponding minimum-loss transfer function (77) is

$$T_m(p) = \frac{d}{b}\left[\frac{gp^2 + ap + b}{p^2 + ep + d}\right] = \frac{E_{L\,\text{Avail.}}}{E_L}, \tag{82}$$

for which

$$\nu_1 = \nu_2 = 1. \tag{83}$$

If (81) satisfies (39) and (40) of Part I, then (81) and the arm immittances in Figs. 11 and 12 will all be Class II functions. The corresponding minimum-loss transfer function is

$$T_m(p) = \frac{1}{g}\left[\frac{gp^2 + ap + b}{p^2 + ep + d}\right] = \frac{E_{L\,\text{Avail.}}}{E_L}, \tag{84}$$

for which

$$\nu_1 = \nu_2 = 1. \tag{85}$$

Let us realize the Class I, *p-r* minimum-loss function

$$T_m(p) = \frac{2.759p^2 + 6.897p + 5.000}{p^2 + 4.000p + 5.000} = \frac{2.759(p + 1.25 + j0.5)(p + 1.25 - j0.5)}{(p + 2 + j1)(p + 2 - j1)}. \tag{86}$$

If in the design, (69) and (72) are satisfied with equal signs, the arm immittances will be minimum real-part functions. Then by (83) and (78), $A_1 = 1$ indicating that the section transfer function (64) will have no flat loss. Using the configuration in Fig. 11, the network is as shown in Fig. 13. It is interesting to note that this circuit requires only eight elements (excluding terminations) in order to realize the transfer function (86), while a conventional, RLC bridged-T would use ten, and a constant-resistance lattice would require sixteen.

If any of the branch immittances used are two pole and zero Class III functions and if (69) and (72) are satisfied with equal signs, either Type D realizations or impractical Brune realizations with unity-coupled coils will be required. The branch immittances would be minimum real-part functions and Type D realization would require eight elements for *each* Class III immittance. If (69) and (72) are satisfied by the inequality signs, then the branch immittances will not be minimum real-part functions. By doing this it will be possible sometimes to obtain Type C realizations (usually 7 elements), or Type B realizations (five elements) for the Class III immittance branches. Of course the price for the reduction in number of elements is additional flat loss.

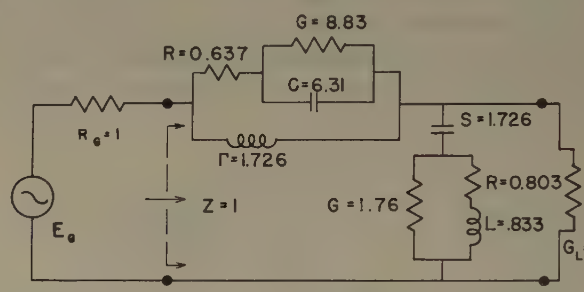


Fig. 13—Example of a constant-resistance ladder design.

In some cases the designer may wish to account for coil and condenser dissipation. If the circuit resulting when (69) and (72) are satisfied with an equal sign cannot be adapted to this, by using the inequality signs (thus increasing the flat loss) it will usually be possible to modify the realization to account for coil and condenser dissipation.

In some cases this method of synthesis will give unequal terminations. It should be noted that if the configuration in Fig. 11 gives a rise in impedance level from input to output, the circuit of Fig. 12 will yield the identical transfer function with a drop in impedance level from input to output. Thus if a chain of sections with unequal terminations are to be connected together, some control can be exerted over the impedance levels at the ends by choosing between the sections of Fig. 11 and Fig. 12, or by using some of each.

When transfer functions are defined as in (75), then the over-all transfer function for a chain of constant-resistance sections is exactly

$$T(p) = [T_1(p)][T_2(p)] \cdots [T_n(p)], \tag{87}$$

where the $T_k(p)$ are the transfer functions of the individual sections. One might at first think that if all of the component sections have minimum-loss transfer functions, then $T(p)$ would also have a minimum-loss transfer function. However note that this is true *only* when the magnitudes of the individual transfer functions all have the value one *at the same frequency* $p = j\omega$. For this reason the designer will sometimes introduce extra loss by realizing a complicated transfer function in a chain of simple sections rather than in one complicated section. The design simplicity obtained will usually be worth the price, however.

It is useful to note that transfer function (82) satisfies (76) with an equal sign at $j\omega=0$ while (84) satisfies (76) with an equal sign for $j\omega=j\infty$. The corresponding no-loss frequencies for other classes of transfer functions with two poles and zeros can be established by the method of Appendix B.

Conclusion

Constant-resistance ladder networks can be used whever conventional, RLC constant-resistance bridged-T networks can be used, i.e., for the realization of any minimum-phase $T(p)$. The constant-resistance ladder has some advantages over the bridged-T, namely: it uses fewer elements; if the bridged-T network requires flat loss the ladder will require less; and since the two ladder branch impedances for each section are not reciprocals of each other as are the bridged-T impedances, the designing of ladders is more flexible. By use of simple formulas it is possible to easily determine the nature of the branch immittance functions so that an appropriate realization technique can be selected.

Appendix B

When computing the attenuation factor A_1 is is often necessary to determine what multiplier B will make

$$T_m(p) = B\,\frac{p^2+ap+b}{p^2+ep+d} = BM(p) \tag{88}$$

a minimum-loss transfer function, i.e., make $|T_m(j\omega)|_{\min.}=1$. If the conditions for (82) or (84) are satisfied, this matter is easily handled as was previously noted. Other cases may be treated as follows:

Let us define $j\omega_m$ as the frequency at which $|T_m(j\omega)|_{\min.}$ occurs. The slope of $|T_m(j\omega)|$ must be zero at $j\omega_m$, hence we may find the proper value of B by checking $|T_m(j\omega)|$ at its frequencies of zero slope. If $j\omega_m=j0$, then

$$B = d/b. \tag{89}$$

If $j\omega_m=j\infty$, then

$$B = 1. \tag{90}$$

To check for $j\omega_m$ at other frequencies one may use the following analysis based on a study of the zeros of $[T_m(p)T_m(-p)-1]$. Form the polynomial in Q:

$$(C^2-4b^2)Q^2+(2DC+4b^2+4d^2)Q+(D^2-4d^2)=0, \tag{91}$$

where $C=2b-a^2$ and $D=e^2-2d$. Find any roots $Q=Q_k$ of (90) which satisfy the condition:

$$x_k = \frac{(Q_kC+D)}{2(Q_k-1)} = \text{(a real, positive number)}. \tag{92}$$

For each value $Q=Q_k$ which satisfies both (91) and (92) there is a zero-slope point of $|T_m(j\omega)|$ at the frequency

$$j\omega_k = j\sqrt{x_k}. \tag{93}$$

If the minimum point is at $j\omega_k$, then

$$B = \sqrt{Q_k}. \tag{94}$$

The proper value of B is the *largest* value obtainable from (89), (90), or (94).

Bifilar Helix for Backward-Wave Oscillators*

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Summary—**A bifilar helix used as the slow wave circuit for the backward-wave oscillator is investigated in this paper. Calculations show that the bifilar helix has a higher impedance than that of the ordinary single-wire helix. The propagating characteristics of the bifilar helix are analyzed and some experimental results of cold measurements are presented.**

It is also shown that with a bifilar helix, a periodic electrostatic field which can be used to focus the electron stream is readily obtained by applying a proper dc potential difference between two helical wires. The same bifilar structure may thus be used both for the propagating and for the focusing circuit. A tube of this kind has been constructed and the experimental results are described.

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Introduction

THIS PAPER DISCUSSES the use of a bifilar helix in a backward-wave oscillator, both as a circuit element and an electrostatic focusing means.

A helix wound of a single wire is generally used in traveling wave tubes as the slow-wave circuit, because of its broadband nature in propagation and its simplicity of construction. A bifilar helix is a helix wound on two wires in parallel (Fig. 1(b)). Each wire is a separate path for the microwave current. The electromagnetic field distribution along the helix depends upon the magnitudes and the phases of the current excited on the two helical wires. In slow-wave structures, it is desirable to have a high impedance, so that greater gain

and output can be attained with a given current flow. It is found that the bifilar helix excited so that currents on the two wires are out of phase has a higher impedance than that of the single-wire helix for the backward-wave field component used in backward-wave oscillators.[1]

In the usual traveling-wave tube, a long and slender electron stream is obtained by using magnetic focusing fields. A recent development in high energy accelerators[2] suggests the possibility of focusing the electron stream by means of periodic fields. We may thus apply different dc potentials to the wires of a bifilar helix such that the potential difference between the wires sets up a periodic electrostatic field which can be utilized to focus the electron stream.

Indeed the same bifilar helix which is advantageous as a slow-wave circuit can also be used to focus the electron stream. A tube using the bifilar helix as the propagating and the focusing structure has been constructed and such focusing has been demonstrated. The tube has operated as a backward wave oscillator using this electrostatic focusing, thus eliminating the need of a magnetic focusing field.

Here we will compare the various modes of a single-wire helix with those of a bifilar helix. We will then describe principles of the periodic electrostatic focusing. Experimental results illustrating electrostatic focusing in a bifilar oscillator structure will be given.

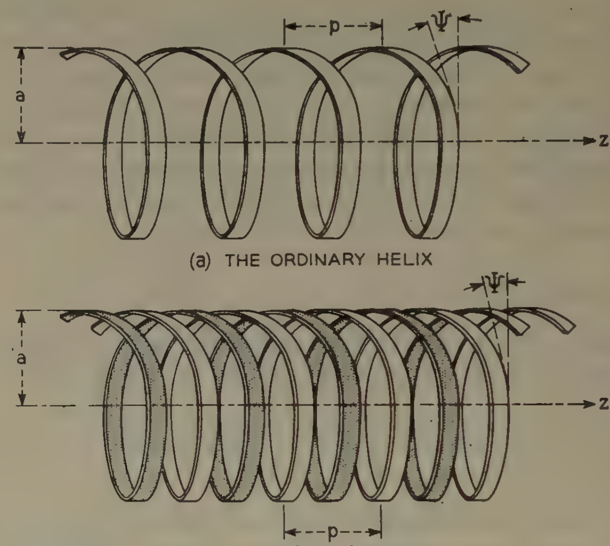
Helix Modes

Various propagating modes of a single-wire helix have been obtained by Sensiper.[3] Those of a bifilar helix are solved in this paper, using the same method. In the tape-helix theory, the helix is assumed to be formed of very thin tape, and the field distribution about this tape-helix as well as the current distribution in the tape, is expressed as a summation of an infinite number of Fourier components such as,

$$\left.\begin{aligned} E^i_{r,z,\theta} &= e^{-j\beta_0 z}\sum_m E^i_{rm,zm,\theta m} I_m\left(\eta_m \frac{r}{a}\right) e^{-jm((2\pi/p)z-\theta)} \\ H^i_{r,z,\theta} &= e^{-j\beta_0 z}\sum_m H^i_{rm,zm,\theta m} I_m\left(\eta_m \frac{r}{a}\right) e^{-jm((2\pi/p)z-\theta)} \\ \kappa_{z,\theta} &= e^{-j\beta_0 z}\sum_m \kappa_{zm,\theta m} e^{-jm((2\pi/p)z-\theta)} \end{aligned}\right\} \qquad (1)$$

$$\eta_m^2 = \beta_m^2 a^2 - k^2 a^2 \qquad (2)$$

where the cylindrical co-ordinates r, z and θ are used. All the quantities with subscript m are associated with the mth Fourier component. Thus, E_{zm}, H_{zm}, and κ_{zm} are respectively the amplitudes of the electric field E, the magnetic field H, and the current density κ of the mth component in the axial direction. Components in the other directions are similarly noted. Superscript i denotes quantities inside the helix. Fields outside the helix can be obtained from (1) by interchanging $I_m(\eta_m(r/a))$ functions with $K_m(\eta_m(r/a))$ functions. $I_m(\eta_m(r/a))$ are respectively modified Bessel functions of mth order of the first and second kinds. m may be any positive or negative integer. The component for which $m=0$ is called the fundamental component. ka is the ratio of the helix circumference to the free space wavelength, a quantity directly proportional to the operating frequency. a is the radius of the helix and p is the pitch as shown in Fig. 1.



(b) THE BI-FILAR HELIX

Fig. 1—(a) Single-wire helix. (b) Bifilar helix.

It is seen from (1) that the fields of the mth component vary angularly as $e^{jm\theta}$ and propagate in the axial direction with a phase constant β_m such that,

$$\beta_m = \beta_0 + m\frac{2\pi}{p}. \qquad (3)$$

This should be so because of the periodicity of the helix. All of these Fourier components are solutions of the wave equation. A summation of them tied together so as to satisfy boundary conditions at the tape forms a mode of propagation. It is evident from (3) that all the components of a common mode must have a common group velocity, but different phase velocities, and some of them may have oppositely directed group and phase velocities.

To solve various modes, the phase constants of the fundamental components are determined from the determinant equation.[3]

$$\sum_m \left[\left(\beta_0^2a^2 - k^2a^2 + k^2a^2\frac{m^2\cot^2\Psi}{\eta_m^2}\right) I_m(\eta_m)K_m(\eta_m) + k^2a^2\cot^2\Psi I_m'(\eta_m)K_m'(\eta_m)\right] D_m = 0 \qquad (4)$$

[1] R. Kompfner and N. T. Williams, "Backward-wave tubes," Proc. I.R.E., vol. 41, pp. 1602–1611; November, 1953.

[2] E. D. Courant, M. S. Livingston, and H. S. Synder, "The strong focusing synchrotron—new energy accelerator," *Phys. Rev.*, vol. 88, pp. 1190–1196; December, 1952.

[3] S. Sensiper, "Electromagnetic Wave Propagation on Helical Conductors," Thesis for Massachusetts Institute of Technology, May, 1951.

where $\cot \Psi = 2\pi\, a/p$. With β_0 known, the phase constants of the other components can be calculated from (3). D_m in (4) is a quantity[3,4] determined by the current distribution assumed over the conducting tape, and for the narrow tape-helix, it is

$$\frac{\sin (m\pi\delta/p)}{m\pi\delta/p}$$

where δ is the width of the tape. The determinant equation is derived based upon the assumption that the current flowing in the direction normal to the helical direction and the electric field along the center line of the tape are zero. The equation can be applied both for the single-wire and the bifilar helices. For a single-wire helix, m includes all the positive and negative integers. For a bifilar helix, m may be limited to the even or odd integers depending upon the conditions of excitation. In general, the solution of (4) is not unique. Each solution is thus identified as a separate mode.

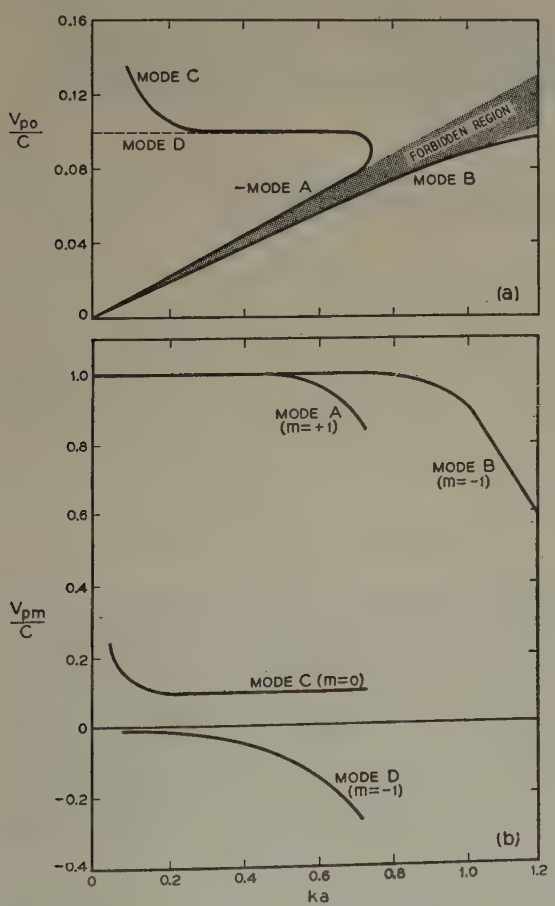


Fig. 2—(a) Phase velocities vs. ka, for the fundamental components of various helix modes ($\cot \Psi = 10$). (b) Phase velocities vs. ka for the dominant components ($\cot \Psi = 10$).

In Fig. 2(a), the phase velocities of the fundamental components are plotted versus ka for various modes. The modes labelled with the letters A, B, and C are respectively the $-h_{t1}''$, h_{t1}' and h_{t0} modes in the Sensiper's analysis.[3] They can be excited with a single-wire helix as well as with a bifilar helix, mode D, however is limited to the bifilar helix. Although the mode D does not contain the fundamental component, its phase velocity is plotted as the dotted line, as computed from (3), in order to compare it with that of the mode C. The phase velocities are plotted in Fig. 2(b) for the $m=+1$ components of the mode A, the $m=-1$ components of the modes B and D, and the $m=0$ component of the mode C. These are generally the dominant components. We mean by the dominant component, the component of the largest amplitude in each mode.

[4] P. K. Tien, "Traveling wave tube helix impedance," Proc. I.R.E., vol. 41, pp. 1617–1623; November, 1953.

Modes A and B are usually denoted as fast waves. At low frequencies, a study of their field patterns shows that their fundamental components are circularly polarized plane waves. The fundamental component of mode A rotates in the direction of the helical wire, and that of mode B the reversed direction as the wave advances. Mode C is the familiar "traveling-wave tube mode." Its fundamental component is generally used in the traveling-wave amplifiers, and the $m=-1$ component, in the backward-wave oscillators. As plotted later the impedance of the $m=-1$ component of this mode is fairly small at low frequencies and increases with ka, becoming comparable to that of the fundamental component at $ka \cong 0.5$.

Mode D is associated with the bifilar helix. Its dominant component which varies angularly as $e^{-j\theta}$, and has oppositely directed group and phase velocities, is therefore of particular interest in the backward-wave oscillator. We may consider this mode as the TEM wave traveling along a coiled two-wire transmission line. As this mode has not been analyzed previously, we will discuss it in detail in the next section.

Solution of Mode D

For the bifilar helix, we are interested in the following two cases: (1) the currents flowing on the two helical wires are equal and in phase, at a constant z plane, and (2) the currents flowing on the two helical wires are equal and in opposite phase at a constant z plane. In the first case, components with $m = \pm 1, \pm 3, \cdots$, are equal to zero, and in the second case, components with $m = 0, \pm 2, \pm 4, \cdots$, are equal to zero. The propagating mode C is excited in the first case and the modes A, B, and D, the second case, as shown in Figs. 3(a) and (b) for $ka < 1.5$.

For simplicity, we will limit our discussion to $ka < 0.8$, for beyond this range the design of a helix becomes impractical. The phase constants of the mode D can easily be solved from (4) as follows: Assuming in the case (2), and $\eta_m \neq 0$ at low frequencies,

$$\frac{I_m'(\eta_m) K_m'(\eta_m)}{I_m(\eta_m) K_m(\eta_m)}$$

can be replaced by $-(m^2+\eta_m^2)/\eta_m^2$ with only few per cent of error. Eq. (4) is then reduced to,

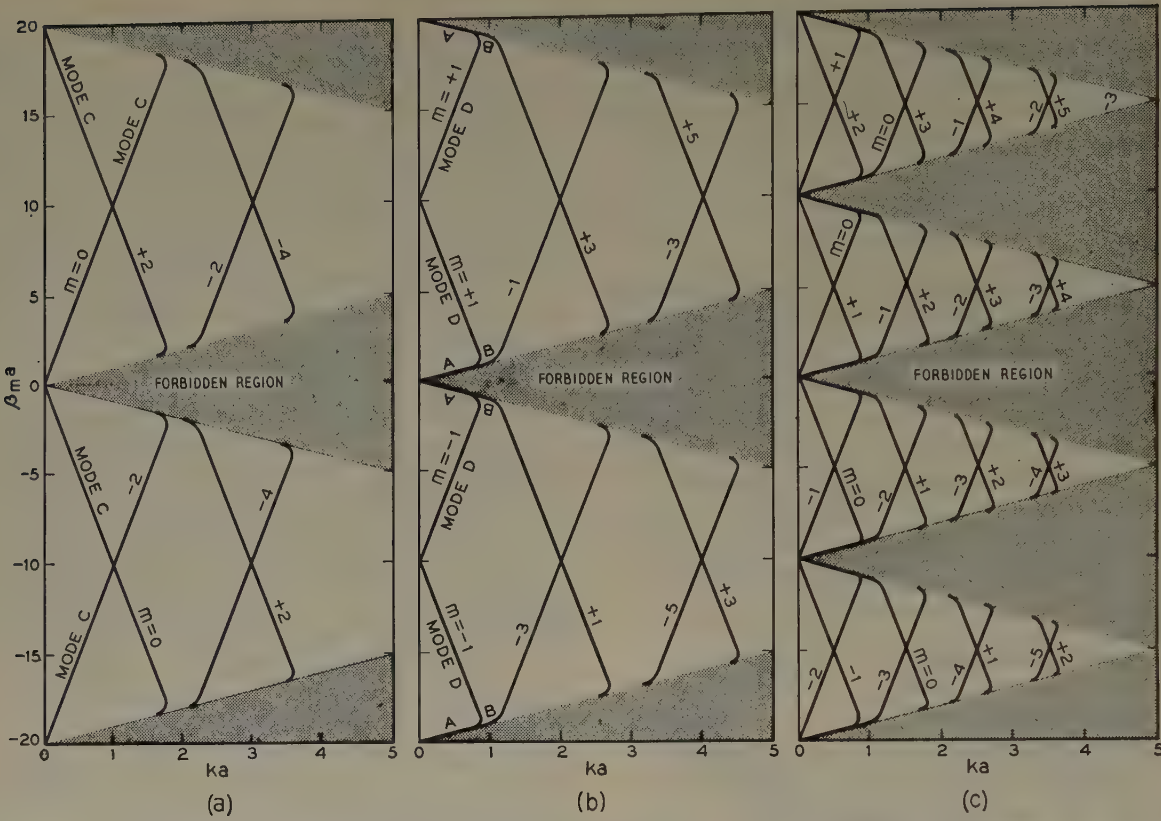


Fig. 3—(a) Propagating characteristics of the bifilar helix (Case 1), cot $\Psi=10$. (b) Propagating characteristics of the bifilar helix (Case 2), cot $\Psi=10$. (c) Propagating characteristics of the single-wire helix, cot $\Psi=10$.

$$\sum_m (\beta_0^2a^2 - k^2a^2 - k^2a^2 \cot^2 \Psi)I_m(\eta_m)K_m(\eta_m)D_m = 0. \quad (5)$$

The above equation consists of an infinite number of terms, all containing a common factor $(\beta_0^2a^2-k^2a^2-k^2a^2 \cot^2 \Psi)$. One of its solutions must then be

$$\beta_0^2a^2 - k^2a^2 - k^2a^2 \cot^2 \Psi = 0$$

or from (2)

$$\eta_0 = ka \cot \Psi. \quad (6)$$

When cot $\Psi \gg 1$, $\beta_0 a \cong \eta_0$; and from (3),

$$\beta_m a = \cot \Psi(m + ka) \quad (7)$$

$$\beta_{-1}a = - \cot \Psi(1 - ka) \quad (8)$$

$$v_{p_{-1}} = - \frac{\omega}{\cot \Psi(1 - ka)} a \quad (9)$$

where v_{pm} is the phase velocity of the mth component and ω is the angular frequency. Comparing the modes D and C, it is seen that the phase velocities of the same Fourier component of both modes coincide with each other beyond the dispersive region of the mode C. For mode D, the dispersive region disappears, simply because the mode D does not contain the fundamental component. This mode has a group velocity close to the velocity of light along the helical direction even at very low frequencies. This indicates that it has a field pattern similar to that of the TEM wave traveling along the coiled two-wire transmission line.

The propagating characteristics of various modes of a bifilar helix are plotted in Fig. 3(a) and (b) for cases (1) and (2) respectively, and are compared with those of a single-wire helix, shown in Fig. 3(c). In these figures, the propagating characteristics (the curves of $\beta_m a$'s versus ka) of each mode are shown as a family of parallel curves with each curve for one component. This should be so according to (3). Curves are given both for the modes traveling in the $+z$ direction and those traveling in the opposite direction. Components are labelled with the numbers, and modes are labelled with letters.

It is interesting to see that for the cases (1) and (2) of the bifilar helix, the forbidden regions repeat at an interval of 2 cot Ψ instead of the interval of cot Ψ for the single-wire helix. The mode D (the case (2) of the bifilar helix) ends at $ka \cong 0.7$ where it meets the mode A, just like the mode C of the single-wire helix, but the mode C of the case (1) of the bifilar helix extends over $ka=1$.

Helix Impedance and Cold Measurements

The helix impedance for traveling-wave amplifiers has been computed in my previous paper.[4] The impedance of the backward-wave components are calculated in a similar way based upon the assumption[3,4] that the magnitude of the current density at the helical cylinder becomes infinitely large in an inverse square root manner as the tape edges are approached, and is zero in the gap between helical conductors, and that the constant phase contour of the current is normal to the edge of the tape.

The impedances are plotted in Fig. 4 both for the $m=-1$ component of the mode D of a bifilar helix and that of the mode C of a single-wire helix. Both helices are assumed in free space. The impedance of the mth component is defined as

$$\frac{E_{zm}^2\ (\text{at } r = a)}{2\beta_m^2 P}$$

where E_{zm} is the axial directed electric field of the mth component evaluated at the helix surface, and P is the total power carried by the circuit (the fundamental and all the harmonics of the mode). It is seen that the impedance of the bifilar helix for the dominant backward component is *much higher* than that of the single-wire helix at low frequencies. This is not surprising, as for the mode C, the fundamental component is stronger than the $m=-1$ component, but for the mode D, the fundamental component is equal to zero.

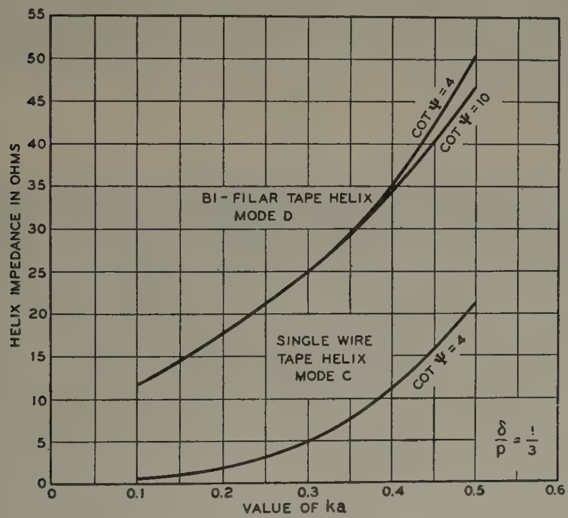


Fig. 4—Helix impedance of the $m=-1$ components for the bifilar and single-wire helices (cot $\Psi=4$, 10).

In order to verify the theory, special equipment has been constructed (Fig. 5), which allows a measuring probe to move accurately in both the z and θ directions. This equipment can be used to measure the standing wave on the helix both along the axial and the helical directions. A bifilar helix of $\frac{1}{2}$ inch in diameter and of pitch of four turns per inch, was wound of 0.030 inch moly. wire as an experimental model. The microwave currents excited on the helix wires were separately adjusted by attenuators and phase shifters. These were carefully adjusted so as to excite only the mode D. Fig. 6(a) shows the standing wave measured along the

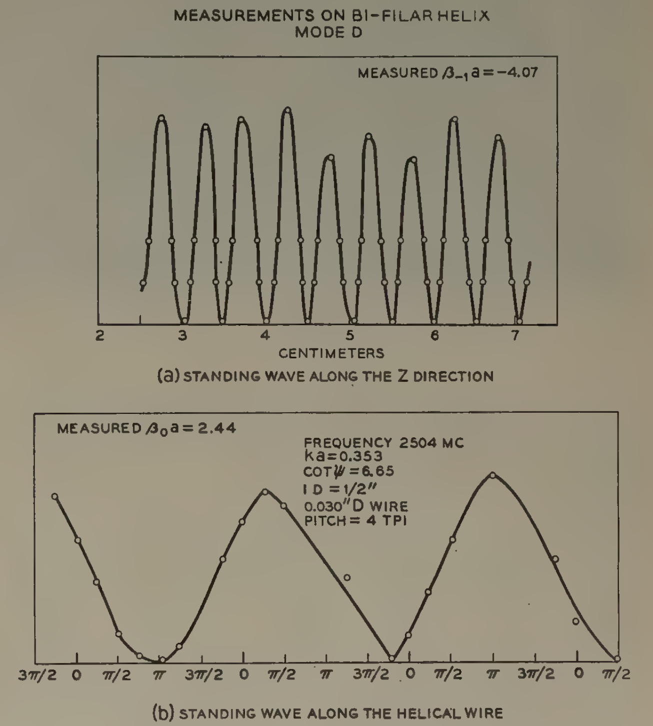


Fig. 6—Standing waves on a bifilar helix. (a) Measured in axial direction (mode D). (b) Measured in helical direction (mode D).

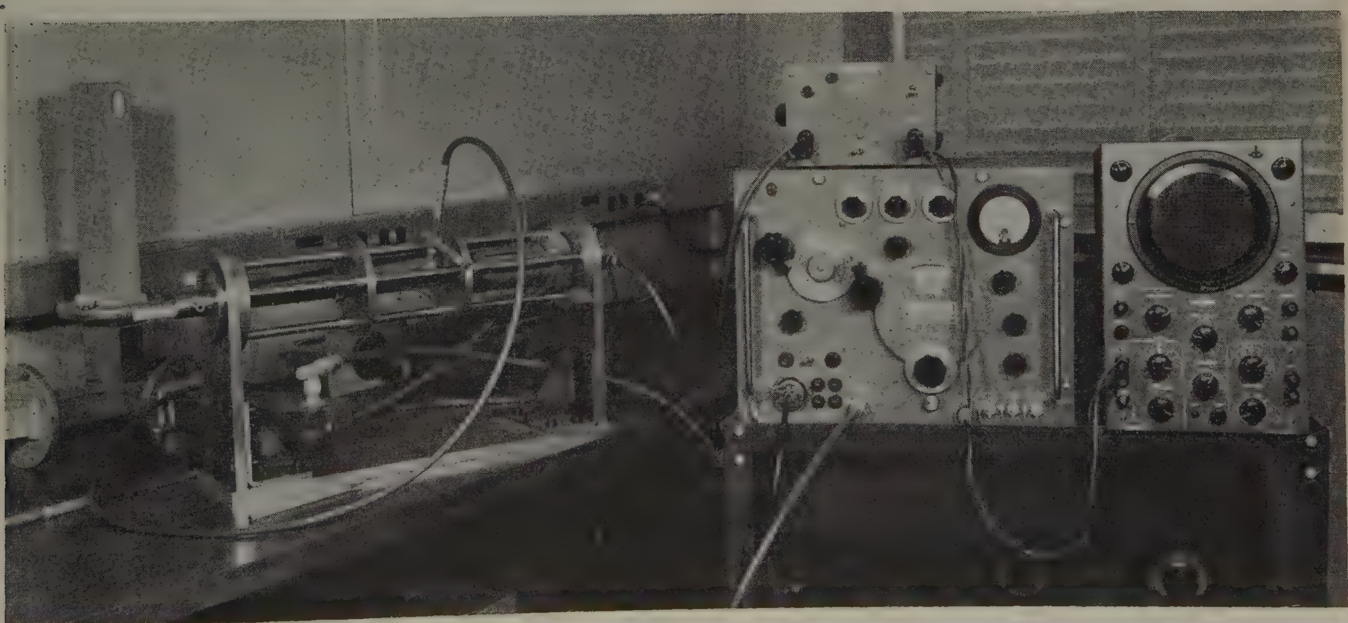


Fig. 5—Equipment used to measure standing waves on the helix.

z direction and Fig. 6(b), that measured along the direction of the helical wire. Both measurements were made at the same frequency, corresponding to $ka = 0.353$. From Fig. 6(b) we are sure that only one mode was propagating, and from Fig. 6(a) we see that $m = -1$ component is the dominant one. If it were not the mode D, the $m = -1$ component would be covered by the stronger $m = 0$ component and would not be observed in the measurements.

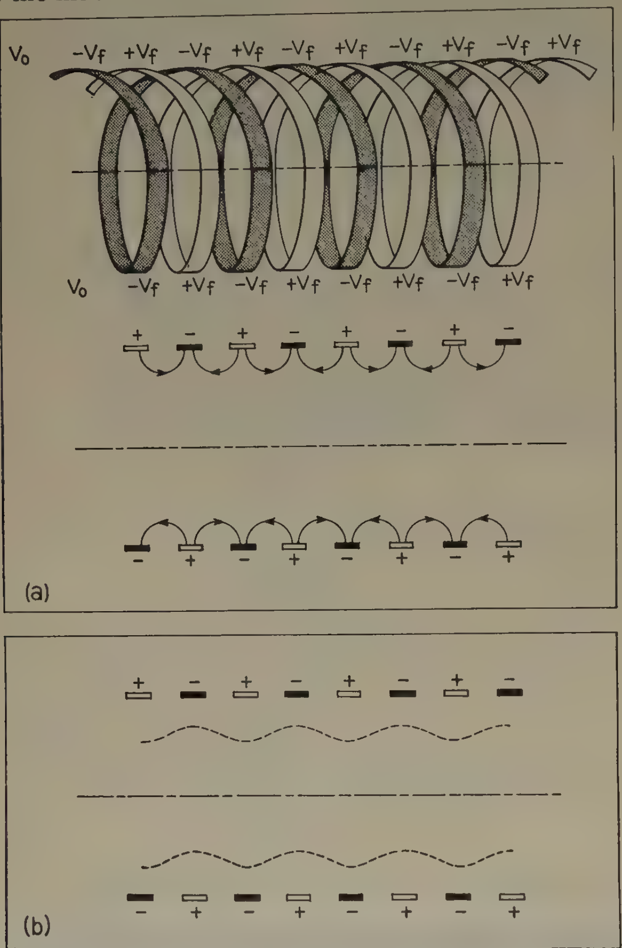


Fig. 7—(a) Periodic electrostatic fields produced by the potential difference between two wires. (b) The radial motion of the boundary electrons due to the periodic electrostatic fields.

Periodic Electrostatic Focusing of the Electron Stream

The periodic electrostatic fields produced by the dc potential difference between two wires of a bifilar helix is shown in Fig. 7(a). It forms a series of alternating converging and diverging lenses of equal strength. Focusing of this kind has been proposed in connection with high energy accelerators. Pierce (unpublished work) has shown that if C is the convergence of lens, and L the spacing between two adjacent lenses, the electron orbit is stable when product CL is less than two.

The principle of focusing may also be explained in another way. The electrostatic potential distribution inside a bifilar tape-helix with potential $V_0 + V_f$ applied to one helical wire and $V_0 - V_f$ to the other, is approximately of the following form

$$V(r, z, \theta) \cong V_0 + \frac{4}{\pi} V_f \frac{\sin\left(\frac{\pi}{2}\left(1 - \frac{2\delta}{p}\right)\right)}{\frac{\pi}{2}\left(1 - \frac{2\delta}{p}\right)} \cdot \sin\left(\frac{2\pi}{p} z - \theta\right) \frac{I_1\left(\frac{2\pi}{p} r\right)}{I_1\left(\frac{2\pi}{p} a\right)} \tag{10}$$

where again, p is the pitch and a, the radius of the helix, and δ is the width of the helical tape. It may be seen that the fields are stronger close to the helix surface and become zero at the helix axis. If we consider a layer of the electrons at the boundary of the stream, the radial electron motion under a perfectly balanced condition is shown in Fig. 7(b). It may be seen that electrons are in a stronger field, when the field is focusing, and in a weaker field, when the field is defocusing. This gives a net focusing force which should be balanced by the space charge forces of the stream. For a solid stream, we may consider electrons at any radius as the boundary electrons of a smaller stream. Because the periodic fields decrease with the radius according to (10), the electron current density must therefore be distributed across the stream such that the space charge fields vary radially in the similar manner, to maintain the balance between the focusing and the space charge forces anywhere inside the stream. Electrons must enter the focusing structure practically without transverse velocities. A quantitative analysis[5] has been made, also indicating that the entering conditions of the stream must be carefully controlled to achieve proper focusing.

A Backward-Wave Oscillator with Periodic Electrostatic Focusing

A backward-wave oscillator using bifilar helix for both focusing and slow-wave circuit is illustrated in Fig. 8. The electron gun consists of a flat cathode of 0.125 inch in diameter, a beam-forming electrode and two electrodes of mesh grid. A 10-inch bifilar helix is wound of 0.030 inch moly. wires and has a mean diameter of 0.186 inch with cot $\Psi = 3.5$.

A typical focusing-curve as directly photographed from the oscilloscope screen is shown in Fig. 9. In this figure, the collector current is plotted versus the potential difference between two helical wires. Ten divisions on the screen correspond 1 milliampere of the collector current. The potential difference between two wires is swept from 0 to 2,300 volts. During the sweeping, the sum of the collector and the helix currents is maintained constant at 1 milliampere. It may be seen that a maximum current transmission of 78 per cent is ob-

[5] P. K. Tien, "Focusing of a long cylindrical electron stream by means of periodic electrostatic fields," (Submitted to *Jour. Appl. Phys.*).

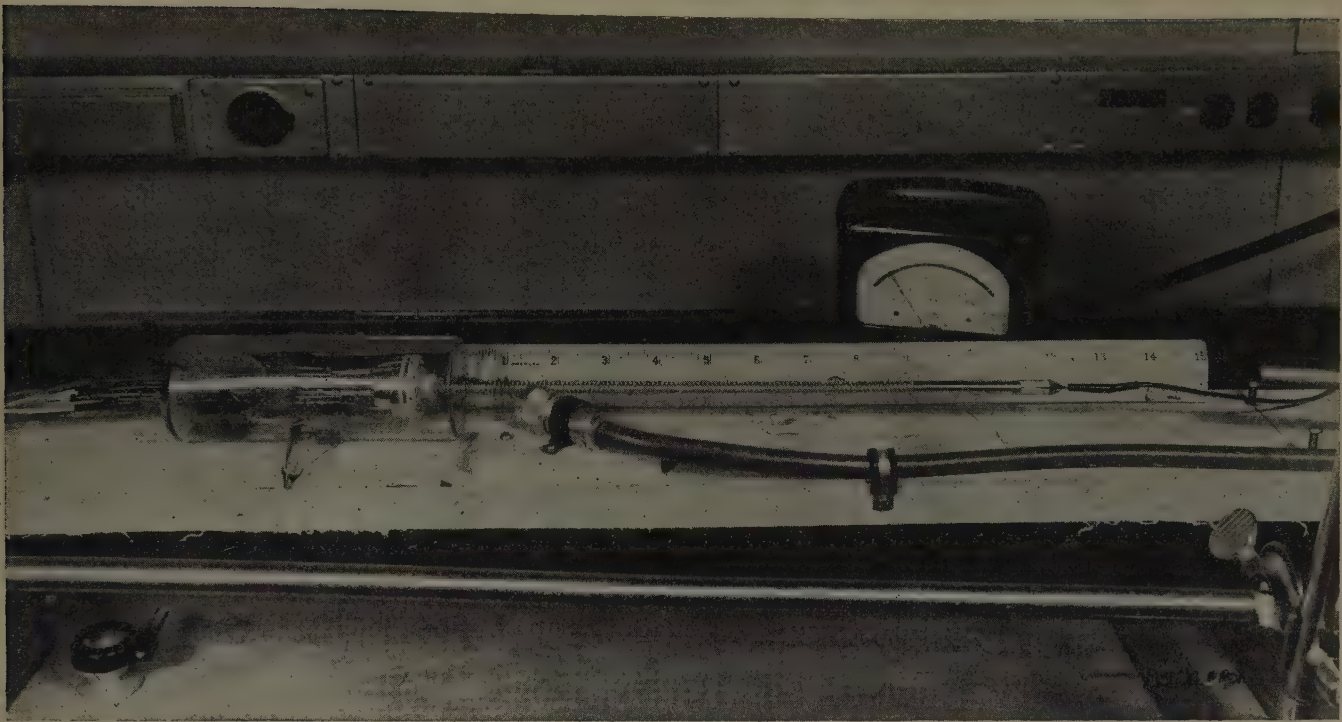


Fig. 8—A backward wave oscillator using a bifilar helix as the focusing and the propagating circuit.

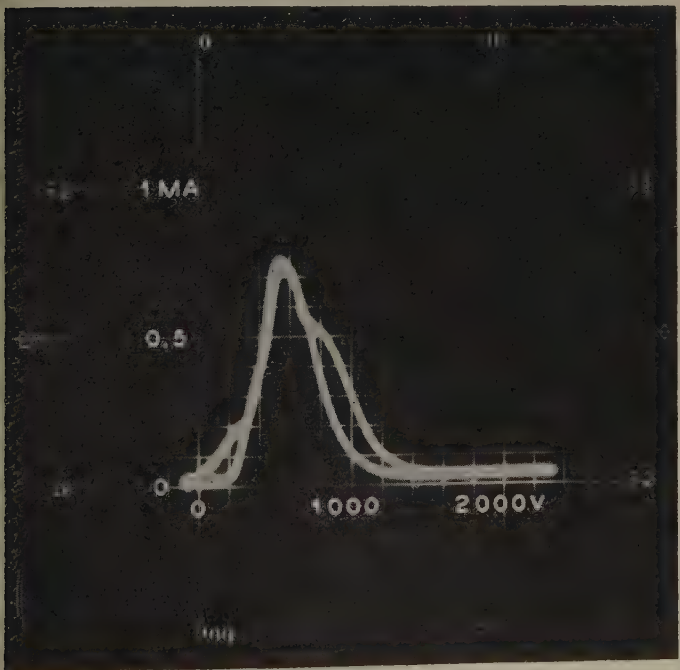


Fig. 9—Collector current vs. dc potential difference between two helical wires.

tained at a potential difference of 480 volts, which is approximately equal to the average beam voltage measured with respect to the cathode.

With periodic electrostatic focusing, oscillations between 3,000 to 3,500 mc have been obtained. The frequency range of oscillation was limited by the amount of the current available from the electron gun, because the starting current required increases with the frequency, and the current transmission becomes poorer when the average potential is higher. The tube was also tested with magnetic focusing. Oscillations between 2,000 to 5,000 mc were observed. The starting current required for the case of the periodic focusing is in general a few times higher than that for magnetic focusing. As the helix is not well terminated at both ends, the measurement on the starting currents does not give correct values for impedance.

Acknowledgment

In this laboratory, the periodic electrostatic focusing has been studied by Pierce based upon the principles of thin lenses. Before the bifilar helix tube was made, C. F. Quate had successfully focused the electron stream by means of periodic quadrupole type of electric fields. Their work has had much influence on the later studies. The experimental tube was made by G. W. Knott, under the supervision of D. J. Brangaccio. The curves were calculated by Miss J. A. Aichele.

Gain Stable Mixers and Amplifiers with Current Feedback*

GAIL E. BOGGS†, MEMBER, IRE

Summary—**Narrow-band radio-frequency amplifiers and mixers may be stabilized by negative feedback without increasing the bandwidth excessively. A couple using current feedback is described. This couple requires only a simple resistive beta circuit and may be designed such that the band-pass characteristic is largely independent of the feedback. Consideration is given to the problem of input impedance and a design procedure is outlined. A discussion of the experimental results concludes the paper.**

Introduction

INVERSE feedback may be used to substantially improve the gain stability of radio-frequency amplifiers and mixers.[1,2,3] The use of voltage feedback results in a reduced output impedance and an increased bandwidth which is often desirable for wide-band applications. When the bandwidth is increased by the application of feedback, as is the case with voltage feedback, the bandwidth becomes dependent upon the zero-feedback voltage gain of the amplifier; the greater the increase in bandwidth the greater is this dependence. This is obviously an undesirable effect since the bandwidth will decrease as the amplifier tubes age even though the center frequency voltage gain remains substantially constant. Narrow-band amplifiers using voltage feedback require an unusually high Q for the tank circuits in order to obtain a relatively narrow bandpass characteristic.

It can be readily shown that current feedback results in an increase in the output impedance of an amplifier and hence in many cases a reduced bandwidth. In amplifiers using pentode tubes the output impedance of the amplifier will generally be sufficiently high without feedback to present negligible loading of the tuned output circuit. Hence, if the bandwidth of a current feedback amplifier is determined by the output circuit, the bandpass characteristic will be essentially independent of both the feedback and the zero-feedback voltage gain.

This paper will be concerned with radio-frequency amplifiers and mixers arranged in a feedback couple with a simple resistive beta circuit. The narrow-band amplifier and mixer couples to be discussed use current feedback and the bandpass characteristic may, by proper design, be largely independent of the feedback.

* Decimal classification: R357.4×R363.1. Original manuscript received by the IRE, March 25, 1953; revised manuscript received, March 11, 1954.

† National Bureau of Standards, Washington 25, D. C.

[1] D. G. Tucker, "Frequency changers and amplifiers with constant gain," Proc. I.R.E., vol. 37, pp. 1324–1327; November, 1949.

[2] G. F. Montgomery, "Intermediate frequency gain stabilization with inverse feedback," Proc. I.R.E., vol. 38, pp. 662–667; June, 1950.

[3] G. E. Boggs, "Improvement in gain stability of the superheterodyne mixer through the application of negative feedback," Proc. I.R.E., vol. 40, pp. 203–207; February, 1952.

Amplifier Theory

Referring to Fig. 1, it is apparent that the feedback voltage for this system is derived from the plate current of the second stage. As a result, the system may be considered as a current feedback couple.

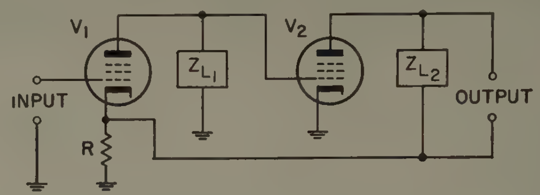


Fig. 1—Schematic of current feedback couple.

The nomenclature to be used in the following brief analysis is given below. It is recognized that many of the quantities listed are properly complex but since we are interested primarily in the center frequency considerations where phase shifts are generally very nearly integral multiples of *pi*, these quantities are listed as reals.

A = voltage gain of couple without feedback
A_1 = voltage gain of first stage
A_2 = voltage gain of second stage
A_f = voltage gain of couple with feedback
A_n = normalized voltage gain of couple
a_1 = Fourier series coefficient proportional to conversion transconductance
b_0 = Fourier series coefficient proportional to the average transconductance
B = return difference
C_1 = grid to cathode capacitance of first stage
C_2 = cathode to ground capacitance of first stage
E_i = maximum value of input voltage
E_0 = maximum value of output voltage
f_0 = center frequency of couple in cycles per second
Δf = bandwidth in cycles per second
g_{m1} = first stage tube transconductance in mhos
g_{m2} = second stage tube transconductance in mhos
g_c = conversion transconductance in mhos
k_0 = constant proportional to the average transconductance
k_1 = constant proportional to the conversion transconductance
Q_k = Q of cathode circuit
Q_L = Q of plate load impedance
R = cathode resistance in ohms
R_L = effective plate load resistance in ohms

$$u=\frac{f}{f_0}-\frac{f_0}{f}\cong\frac{\Delta f}{f_0}$$

$x = Q_L u$
y_i = input admittance in mhos
Z_{L1} = plate load impedance of first stage in ohms
Z_{L2} = plate load impedance of second stage in ohms
ω_1 = angular frequency of signal voltage
ω_2 = angular frequency of oscillator voltage.

Fig. 2 gives the equivalent circuit of the couple. The grid-to-cathode and cathode-to-ground capacitors are shown since input admittance considerations will be considered later. For gain determinations, these capacities will be neglected and the equivalent circuit reduced to three node pairs.

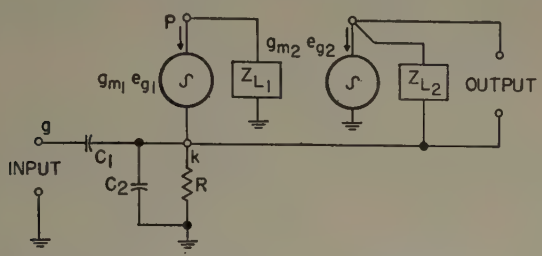


Fig. 2—Equivalent circuit of current feedback couple.

From the three nodal equations

$$E_0 = \frac{g_{m1}g_{m2}Z_{L1}Z_{L2}E_i}{1 + g_{m1}R + g_{m1}g_{m2}Z_{L1}R}. \tag{1}$$

If local feedback on the first stage is neglected ($g_{m1}R \ll 1$) then

$$E_0 \cong \frac{g_{m1}g_{m2}Z_{L1}Z_{L2}E_i}{1 + g_{m1}g_{m2}Z_{L1}R}. \tag{2}$$

Now by definition, the gain with feedback is,

$A_f = \dfrac{E_0}{E_i}$ and the zero feedback gain (A)

$$A \cong g_{m1}g_{m2}Z_{L1}Z_{L2}.$$

Substituting in (2)

$$A_f \cong \frac{A}{1 + \dfrac{AR}{Z_{L2}}}. \tag{3}$$

Now by definition,

$$B = \frac{AR}{Z_{L2}} \tag{4}$$

and

$$A_f \cong \frac{A}{1 + B}, \tag{5}$$

which is a conventional form of feedback equation.

Bandwidth Considerations

In many applications of this type of amplifier it is desirable to have the bandwidth independent of the feedback. Therefore, the effect of the interstage tank circuit is considered briefly. For simplicity consider a resistive load impedance R_L on the second stage. The gain, A_f and A must be considered to be complex for the purpose of this discussion and the subscript 0 denotes center frequency gain.

From (3), it may be shown that

$$A_f = \frac{A_{f0}}{1 + \dfrac{jQu}{1 + \dfrac{A_0 R}{R_L}}}. \tag{6}$$

By inspection of (6) it is apparent that the effective Q of the interstage circuit is reduced by the feedback factor $1+B$, as is the case with a single stage, voltage feedback amplifier. If the resistive load R_L is now placed on the first stage and the complex impedance Z_{L2} on the second stage, (3) is still valid but A is now $g_{m1}g_{m2}R_LZ_{L2}$. Following a similar procedure yields

$$A_f = \frac{A_{f0}}{1 + jQu}. \tag{7}$$

Obviously the bandpass characteristic is unaffected by the application of feedback provided pentode tubes are used. For amplifiers using similar tank circuits and feedback factors of ten or greater, the bandpass characteristic is essentially that of the output circuit. The bandwidth of such a system will be largely independent of the zero-feedback voltage gain until the gain and hence the feedback factor, have fallen to a level such that the interstage network begins to make an appreciable contribution to the over-all selectivity. In addition, this system permits the use of the feedback couple, with many of the attendant advantages, without the necessity of using a critical reactive beta circuit. Obviously, if the interstage Q is less than that of the output circuit, the bandwidth will remain independent of voltage gain for lower values of $1+B$.

Input Impedance

It is well known[4] that with a cathode follower, the real part of the input admittance may be negative. With couples requiring a large value of cathode resistance or operating at frequencies of approximately 10 mc or more, the negative input admittance may result in oscillation at or near the signal frequency when the couple is operated with a high impedance input. This discussion will be limited to input admittance effects due to the unbypassed cathode of the first stage.

Referring again to Fig. 2, the nodal equations may be solved for the real part of the input admittance y_i. Substituting for the plate load impedances $Z_{L1} = Z_{L2} = R_L/1 + jx$, and assuming that $g_mZ_L \gg 1$, gives

$$R_e y_i \cong \frac{\omega^2 C_1^2 R[x^2 + 1] - \omega C_1 B[x + \omega C_2 R]}{x^2 - 2\omega(C_1 + C_2)RBx + (1 + B)^2}. \tag{8}$$

[4] F. D. Clapp, "Some aspects of cathode-follower design at radio frequencies," Proc. I.R.E., vol. 37, pp. 932–937; August, 1949.

Inspection of (8) shows that a negative input conductance may occur if the second term in the numerator exceeds the first term. Closer examination reveals two regions in which this may occur, one at the center frequency ($x=0$) if $B>C_1/C_2$ and the other at a slightly higher frequency (x positive). Fortunately, the cathode to ground capacitance may be reduced at center frequency by shunt resonating the cathode. Then the input conductance will always be slightly positive at the center frequency. Even under this condition, a high frequency couple could conceivably oscillate at a higher frequency if the driving impedance is of the correct form.

Mixer Couple

If the first tube of the couple is replaced by a mixer such as the 6SA7, the difference frequency voltage may be used as negative feedback to stabilize the mixer and amplifier operation.

Following the procedure of an earlier paper,[3] the output voltage of the mixer couple may be shown to be

$$E_0 = \frac{a_1 g_{m2} E_i Z_{L1} Z_{L2}}{2(1 + b_0 g_{m2} R Z_L)} \cdot \tag{9}$$

For simplification let

$$b_0 = k_0 \frac{g_{m1}}{\pi} \quad \text{and} \quad a_1 = k_1 \frac{g_{m1}}{\pi}$$

where k_0 and k_1 are constants determined by the switching function of the mixer tube selected.

Writing (9) in terms of gain and substituting for b_0 and a_1

$$A_f = \frac{A}{1 + \dfrac{2k_0 RA}{k_1 Z_{L2}}} \cdot \tag{10}$$

The return difference may be defined by

$$B = \frac{2k_0 RA}{k_1 Z_{L2}} \tag{11}$$

and (11) reduces to

$$A_f = \frac{A}{1+B} \cdot$$

Obviously, for large values of B, the gain is relatively independent of the transconductance of either tube. It should be noted that k_0 and k_1 are subject to variation due to changes in the shape of the switching function which may result from a change in oscillator voltage as well as other causes. While this may be considered as a limiting factor for stability improvement, experimental results indicate that for pentagrid tubes, these constants tend to change together. With a high degree of feedback, inspection of (10) indicates little change in A_f with changes in the value of the constants, provided that the changes of k_0 and k_1 are in the same direction and of like percentage. Typical values of the constants for a 6SB7-Y are $k_0=1.2$ and $k_1=1.75$. The author has observed that larger variations occur in the values of the constants of pentode mixers.

Since the conversion transconductance is less than the amplifier transconductance, a larger cathode resistor will be required to provide the degree of feedback obtained with an amplifier couple. At the signal frequency, the input admittance may be negative due to the unbypassed cathode. In fact, in some instances, Colpitts oscillation may take place in the signal circuit. The deleterious effects of a negative input impedance at the signal frequency can usually be minimized by a suitable filter on the mixer cathode or by a low impedance signal-grid circuit.

Design Procedure

For the procedure which follows it will be assumed that the load impedances are identical. The design engineer may use this procedure with suitable modification for both the mixer and amplifier couples. Other variations may be readily obtained to fit a particular case. It will be further assumed that the feedback factor $1+B$, is large enough to permit the assumption that the bandwidth is determined only by the output circuit. No consideration is given to the problem of input admittance.

Given: Δf, f_0, A_f, $1+B$, g_{m2} and g_{m1} or g_c.

1. Calculate $A=(1+B)A_f$.
2. For mixer operation determine values of k_0 and k_1 for the tube under consideration. These constants may be determined by graphical integration of the switching function.[3] If this is not expedient, assume $k_1=1.5$ and $k_0=1.0$ and continue solution; suitable corrections must then be made experimentally when the unit is under test.
3. Calculate the gain of each stage:
   for amplifiers $A_1=A_2=\sqrt{A}$,
   for mixers let $k=g_{m2}/g_c$
   $$A_2 = \sqrt{kA},$$
   and
   $$A_1 = \frac{A}{A_2} \cdot$$
4. Calculate $R_L=A_2/g_{m2}$.
5. Calculate $Q=f_0/\Delta f$.
6. Calculate $C=Q/\omega R_L$.
7. Calculate R;
   for amplifiers $R \cong BR_L/A$;
   for mixers $R=k_1 BR_L/2k_0 A$.

Alignment Procedure

To remove feedback for alignment and the measurement of gain, the feedback line should be broken and the cathode of the first tube adequately bypassed. The output circuit must now be returned to ground through a value of resistance approximately equal to R. The plate circuits may now be aligned to the correct frequency and the gain measured. When feedback is reapplied to

the amplifier, the output circuit only should be readjusted for maximum at the same alignment frequency. If the peak of bandpass response does not occur at alignment frequency, output circuit should be adjusted for this condition. A slight response shift occurs in some units due to extraneous phase shift of feedback voltage.

It is worthy of mention that steps must be taken to minimize regeneration since an amount which normally would be tolerated may result in violent performance deviations in a feedback couple.

Experimental Results

For amplifiers operating at low frequencies and with relatively small values of $1+B$ it is often possible to replace the interstage tank circuit with a resistor without encountering excessive phase shift. A 450 kc amplifier, with each plate circuit tuned, was tested experimentally with the results shown in Fig. 3. In this case the gain required resulted in a feedback factor of only 23 db. The bandwidth under each condition of operation is noted on the curve. The type of amplifier described in this paper has been operated successfully at 15 mc and can probably be used at higher frequencies if suitable precautions are taken to restrict extraneous phase shift of the feedback voltage. Obviously, the input admittance problems discussed earlier must be given full consideration at frequencies in the order of 50 mc.

The use of this system for mixer couples results in operation very similar to that obtained with the amplifiers just described. The stability improvement of mixer couples is, in general, not as great as that obtained in amplifiers due to the variation in the constants k_0 and k_1 described earlier. Satisfactory operation has been obtained with a 6AK5 pentode mixer operating at a 50-mc signal frequency and a 5-mc IF frequency. The IF stage of this couple was also a 6AK5.

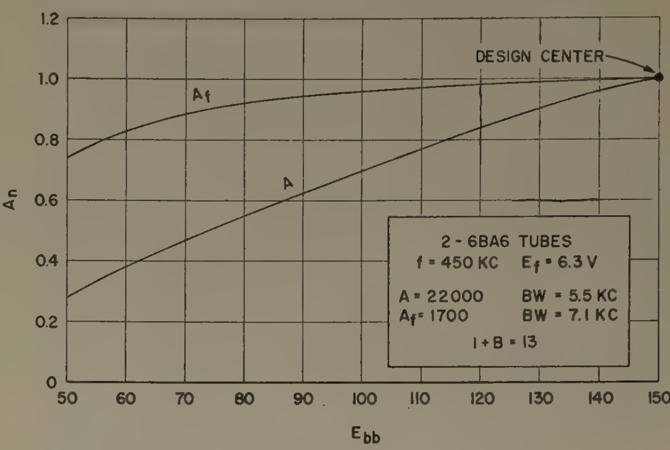


Fig. 3—Gain stability of 450 kc amplifier couple.

Conclusion

It has been shown that current feedback may be used to stabilize radio-frequency amplifier and mixer couples without the use of a reactive beta circuit. The use of this system will permit the design of a narrow-band amplifier whose bandwidth is essentially independent of the feedback.

Since the output circuit is essentially unaffected by the application of feedback to an amplifier using pentode tubes, a complex filter designed to provide the desired band pass characteristic may be readily employed. It should be noted that the impedance of any paths from the output circuit to ground must be much larger than the beta circuit impedance.

Electron Trajectories in Strip Beams Constrained by a Magnetic Field*

J. D. LAWSON†

Summary—**Calculations are made of the electron trajectories in infinite strip beams constrained to move in a magnetic field parallel to the direction of flow. Space charge is taken into account, but nevertheless the analysis is only valid within certain well-defined limits. Different types of solution are obtained according to whether the beam is launched from a source outside or inside the magnetic field. When the gun is outside the field the type of flow also depends on whether the current is greater or less than a certain value. In all cases the beamwidth varies periodically along the length of the beam, and expressions for the periodicity and the ratio of maximum to minimum width are obtained. Both fundamental and practical limitations are then considered, and their effects on the electron flow discussed qualitatively. The importance of conditions outside the beam boundary is emphasized.**

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Introduction

The BEHAVIOR of electron beams possessing circular symmetry and moving in an axial magnetic field has been studied by several authors. A simple analysis which shows clearly the essential nature of the flow has been given by Pierce,[1] and more general treatments which are necessarily more complicated have been given by Convert[2] and by Wang.[3] The behavior of strip beams in a magnetic field has not been so fully

[1] J. R. Pierce, "The Theory and Design of Electron Beams," Van Nostrand Book Co., New York, N. Y., Chap. IX; 1949.

[2] G. Convert, "Étude de la focalisation maguétique de faisceaux electroniques cylindriques," *Annales de Radioelectricité IV*, p. 279; October, 1949.

[3] C. C. Wang, "Electron beams in axially symmetrical electric and magnetic fields," Proc. I.R.E., vol. 38, p. 135; February, 1950.

investigated, though several special cases have been considered in the past.[1,4,5]

In this paper expressions are obtained for the trajectories of the electrons in strip beams constrained by a magnetic field parallel to the direction of flow. The rationalized MKS system of units (listed at the end of the paper) is used, and the terminology and notation of Pierce[1] are followed as closely as possible. In order to avoid mathematical complexity, only simple field configurations are considered, and assumptions are made which are not valid for very high fields or currents. Nevertheless the different types of flow are clearly distinguished, and any practical design is likely to lie in the range where the analysis is valid. The behavior of strip beams is found to be similar in many respects to that of circular beams, though there are some interesting differences.

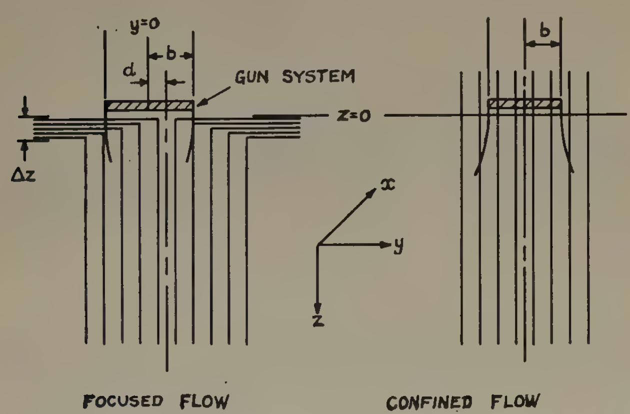

Fig. 1.—Diagrammatic representation of the gun- and magnetic-field configuration for focused and confined flow systems.

Following Pierce, we distinguish two kinds of flow, magnetically-focused flow and magnetically-confined flow. In the former the electron beam is generated outside the magnetic field, which it enters at the $z=0$ plane; in the latter the cathode from which the beam originates is in the magnetic field. The two types of field configuration are shown in Fig. 1 together with the coordinate system and dimensional parameters. A somewhat unnatural field shape near the $z=0$ plane is assumed in the case of magnetically-focused flow, but the nature of the flow will not be essentially different if the beam enters the field more gradually.

Calculation of Trajectories

In this section the trajectory of any electron in the beam is calculated in terms of its position in the $z=0$ plane. We first find the relation between electron current I and electric field in the y direction E_y, we then calculate the impulse in the x direction which the electron receives from the y component of magnetic field near the $z=0$ plane (this is zero in the case of confined flow), and finally we calculate the electron trajectories in the x and y directions due to the crossed electric and magnetic fields E_y and B_z. We note that E_y is a function of z, but is constant for a given electron, provided that the electron trajectories do not cross. With focused flow the trajectories can cross, the condition for this is found and its effect considered.

The assumptions made in the analysis are listed below, and numbered for future reference.

A1. The beam is infinite in the x direction, or possesses circular symmetry about an axis $y=a$ where $a \gg b$. (End effects in finite beams are discussed later.

A2. At the $z=0$ plane the electron density in the beam is uniform, and the electrons are moving in parallel paths with equal velocities given by

$$\dot{z} = \sqrt{2\eta V_0}. \quad (1)$$

A3. Any components of velocity acquired in the x and y directions are small compared with $\dot{z}$. The condition for this to be true is found later (inequality (11)).

A4. For positive values of z the electric field both inside and outside the beam is in the y direction except at the $y=0$ plane where it vanishes. (This assumption is discussed in detail later.)

A5. The velocity of the electrons is small compared with that of light.

To find the relation between I and E_y we use (1) and also the equations

$$I = -2b\rho\dot{z} \quad (2)$$

$$E_y = \rho y/\epsilon \text{ (Gauss' theorem)}, \quad (3)$$

whence by elimination of ρ and $\dot{z}$

$$E_y = -Iy/2\sqrt{2}\,\epsilon\eta^{1/2}bV_0^{1/2}. \quad (4)$$

In the case of the focused flow the electrons will acquire a velocity in the x direction when passing between the $z=0$ and $z=\Delta z$ planes. This arises from the impulse imparted by the y component of magnetic field. Assuming that this acquired velocity is small compared with $\dot{z}$, we may write

$$\dot{x}_1 = \eta\int_0^{\Delta z/\dot{z}} B_y\dot{z}dt = \eta\int_0^{\Delta z} B_y d^z. \quad (5)$$

From the configuration of the lines of force as shown in Fig. 1 we see that the right-hand side of (5) can be expressed in terms of B_z so that

$$\dot{x}_1 = -\eta B_z(y_1 - d). \quad (6a)$$

When the flow is confined, the electrons cross no lines of force when they enter the magnetic field, so that the equation corresponding to (6a) is

$$\dot{x}_1 = 0. \quad (6b)$$

In both focused and confined flow the electrons, after crossing the $z=\Delta z$ plane, are subject to the effect of crossed electric and magnetic fields B_z and E_y. The equation of motion in such fields are

$$\ddot{x} = -\eta B_z\dot{y}$$
$$\ddot{y} = \eta B_z\dot{x} - \eta E_y. \quad (7)$$

[4] A. V. Haeff, "Space-change effects in electron beams," Proc. I.R.E., vol. 27, p. 586; September, 1939.
[5] L. Brillouin, "A theorem of Larmor and its importance for electrons in magnetic fields," *Phys. Rev.*, vol. 67, p. 260; April, 1945.

The solution of these equations for initial conditions $x=x_1$, $y=y_1$, $\dot{x}=\dot{x}_1$, may be written (dropping the subscripts for B and E):

$$\begin{aligned} x &= x_1 + \frac{Et}{B} + \left\{\frac{\dot{x}_1}{\eta B} - \frac{E}{\eta B^2}\right\} \sin \eta Bt \\ y &= y_1 + \left\{\frac{\dot{x}_1}{\eta B} - \frac{E}{\eta B^2}\right\} (1 - \cos \eta Bt). \end{aligned} \tag{8}$$

Substituting for E and $\dot{x}_1$ from (4) and (6), and substituting I_0 as defined in the list of symbols, we get two sets of trajectories for the focused and confined flow respectively.

$$\begin{aligned} x &= x_1 + \{y(I/I_0 - 1) + d\} \sin \eta Bt \\ &\quad - (I/I_0)\eta B y_1 t \\ y &= y_1 + \{y_1(I/I_0 - 1) + d\}(1 - \cos \eta Bt) \\ z &= \sqrt{2\eta V_0}\, t, \end{aligned} \quad \text{(9a, focused flow, provided that the trajectories do not cross)}$$

$$\begin{aligned} x &= x_1 + y_1(I/I_0)(\eta Bt - \sin \eta Bt) \\ y &= y_1\{(I/I_0 + 1) - (I/I_0) \cos \eta Bt\} \\ z &= \sqrt{2\eta V_0}\, t. \end{aligned} \quad \text{(9b, confined flow)}$$

From these we see that in either case the beam edge fluctuates sinusoidally with z. In the confined flow the beam always diverges on entering the field, whereas a focused beam will converge or diverge according to whether I is less or greater than I_0. For focused flow the beam edges are parallel planes when $I=I_0$ and $d=0$. Flow such that the beam contour is independent of z is known as "Brillouin Flow";[5] in beams with circular cross section the current density for such flow is only half that for strip beams with corresponding values of B and V_0.

These equations are only valid if $\dot{x}$ and $\dot{y}$ are small compared with $\dot{z}$ (assumption *A3*). As a rough criterion we may specify that $\dot{z}$ for the outer electrons should be much greater than the mean value of $\dot{x}$. From (9a) or (9b) we get

$$(\dot{x}/\dot{z})_{\text{mean}} = Bb\eta^{1/2}I/\sqrt{2}\, V_0^{1/2}I_0; \tag{10}$$

substituting for I_0 and η the criterion becomes

$$I \ll 6.2V_0B. \tag{11}$$

This will apply in most cases of practical interest.

From (9) we see that the beam contour repeats in a distance in the z direction given by

$$\lambda = 2\pi\dot{z}/\eta B = 2\sqrt{2}\,\pi V_0^{1/2}/\eta^{1/2}B. \tag{12}$$

This is independent of current (unless the trajectories cross, see below) and less than the corresponding figure for cylindrical beams. We can also find the ratio of the maximum (or minimum) beam width to the width at entry to the field by putting $\eta Bt=\pi$ in the expressions for y in (9). This yields (for all values of d):

$$S = 2I/I_0 - 1 \text{ for focused flow } (I > I_0/2) \tag{13a}$$

$$S = 2I/I_0 + 1 \text{ for confined flow.} \tag{13b}$$

For $I<I_0/2$ the value of S in (13a) becomes negative. An examination of the trajectory equations shows that when this condition holds the electrons all pass through a line parallel to the x axis given by

$$\begin{aligned} v_0 &= dI_0/(I_0 - I) \\ z_0 &= \{\sqrt{2}\, V_0^{1/2}/(\eta^{1/2}B)\} \cos^{-1} \{I/(I_0 - I)\}. \end{aligned} \tag{14}$$

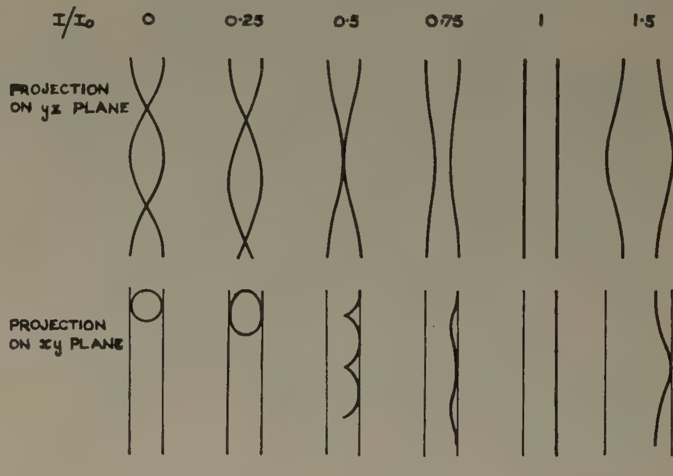

FIG. 2A

Fig. 2(a)—Trajectories for focused flow with $d=0$ and various values of I/I_0. Top diagrams may be regarded either as cross sections of the beam viewed in the x direction, or as the projection on the yz plane of the trajectories of electrons launched from the edge of the gun. Second row diagrams show the trajectories of a typical edge electron projected on a plane perpendicular to direction of flow.

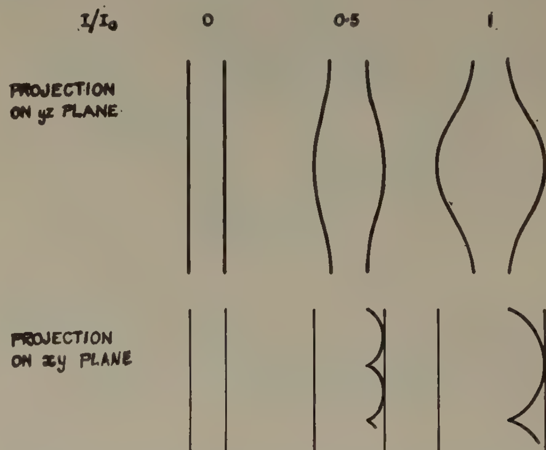

Fig. 2(b)—Diagram as Fig. 2(a) but for confined flow.

For values of z greater than z_0 as given by (14) the trajectory equations (9) no longer hold, since the direction of the electric field acting on the electrons changes sign as they pass through the line (y_0, z_0). The trajectories for values of z greater than z_0 are just reflections through the point where they cross the line (y_0, z_0) of the trajectories between $z=0$ and $z=z_0$. The value of λ is therefore

$$\lambda(I < I_0/2) = [2\sqrt{2}\, V_0^{1/2} \cos^{-1} \{I/(I_0 - I)\}]/\eta^{1/2}B. \tag{15}$$

Figs. 2(a) and 2(b) show the different types of flow discussed here. Projections on the yz and xy planes of the trajectories of electrons at the edge of the beam for different values of I/I_0 are shown. Fig. 2a refers to focused flow (with $d=0$), and Fig. 2b to confined flow.

The analysis of this section must be used with great care. The range of validity of the trajectory equations has already been discussed, but there are also physical limitations due to the impossibility of realizing the conditions assumed at entry to the magnetic field. These are considered in the next section.

Examination of Some Assumptions

We have assumed above (*A2*) that electrons enter the magnetic field in a parallel beam of uniform density, with a uniform velocity distribution of the electrons across it. Except for the small effect of thermal velocities this could be achieved by using a gun consisting of a plane temperature-limited cathode with a grid at potential V_0 very close to it (to avoid end effects and beam spread) and very close to the $z=0$ plane. We have also assumed (*A4*) that after the beam leaves the grid the electric field is in the y direction everywhere except in the $y=0$ plane where it vanishes. At the grid, however there can be no y component of electric field since the grid is at a uniform potential, and, as we shall show, the electric field configuration in the beam near the $z=0$ plane depends on conditions outside the beam boundaries. As a specific example we consider two conducting planes held at a potential V' with respect to the grid, distant g from the beam center as shown in Fig. 3,

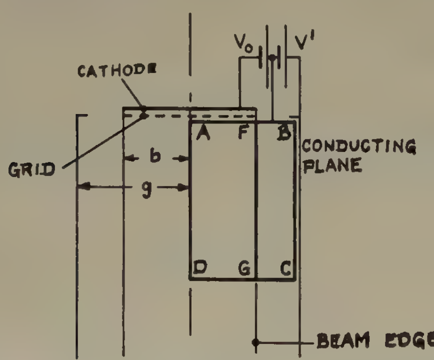

Fig. 3—Diagram showing the gun system and paths of integration used in discussing the validity of the assumption *A4*.

and we assume that the electron flow is such that the beam boundaries are parallel planes. We further assume that the departure from our postulated conditions near the grid has little effect on flow at a large distance from the grid. By this perturbation method we shall find the condition for validity of assumption *A4*.

Consider the path *ABCD* shown in Fig. 3. The integral of the electric field round this must be zero. Integrating round the sides in turn we have

$$-V' + 0 - \int_C^G E_y ds - \int_G^D E_y ds - \int_D^A E_z ds = 0. \quad (16)$$

Now from (3)

$$E_y = \rho y/\epsilon \quad (y \leq b)$$
$$= \rho b/\epsilon \quad (y > b).$$

Whence by substitution for E_y in (16)

$$V' + (g - b/2)\rho b/\epsilon + \int_D^A E_z ds = 0. \quad (17)$$

Similarly for the path *FBCG* we have

$$V' + g\rho b/\epsilon + \int_G^F E_z ds = 0. \quad (18)$$

By comparing these two equations we see that E_z cannot be zero both at the center and at the edge of the beam as we have postulated. Some electrons will therefore be accelerated or decelerated according to the value of the potential V'. For this acceleration or deceleration to be unimportant we must have both

$$\left| \int_D^A E_z ds \right| \ll |V_0| \quad \text{and} \quad \left| \int_G^F E_z ds \right| \ll |V_0|. \quad (19)$$

Now since both these integrals are small, their difference must be small, so that from (17) and (18) we have

$$-b^2\rho/2\epsilon \ll V_0, \quad (20)$$

whence from (2) and (4)

$$V_0 \gg Ib/4\sqrt{2}\, V_0^{1/2}\epsilon\eta^{1/2} \quad (21)$$

or, substituting for ϵ, η, and $I/V_0^{3/2}$,

$$k \ll 21 \times 10^{-6}/b. \quad (22)$$

To ensure therefore that none of the electrons in the beam is appreciably accelerated or decelerated by the electric field in the z direction this inequality must be satisfied, and also V' must be chosen in such a way that inequality (19) is satisfied. This should be possible in a useful number of practical cases.

A more detailed and general analysis of this effect for strip beams is given by Haeff.[4]

If g is large, V' must be large to satisfy condition (19), whereas for systems with circular symmetry this need not be so, since the electric field falls off as $1/r$ away from the beam.

In view of the considerations of this section, and other limitations noted earlier, it must be emphasized that the trajectory equation must be used with extreme caution. The validity of the inequalities (11), (19) and (22) must be carefully checked, and when focused flow is being considered the sign of $2I/I_0 - I$ must be determined to see whether the trajectories cross.

Some Practical Considerations

Besides the theoretical limitations already discussed there will also be practical difficulties in realizing the types of flow considered. It is not possible to produce a sudden transition into the magnetic field, nor can beams with strictly parallel electron paths and uniform charge density be generated. Especially unrealistic is the case with magnetically-focused flow when the trajectories cross. In any practical system, nonuniformity of the beam density in the y direction and the gradual

transition into the magnetic field will blur out the line (y_0, z_0) into a finite region, so that successive focuses become increasingly blurred as z increases; the motion will become more and more disordered and the y dimension of the beam will increase.

It is interesting also to inquire what happens at the ends of beams having finite extension in the x direction. It does not appear possible to arrange the end conditions in such a way that the electrons turn round and travel back along the length of the beam in an orderly fashion. The electrons will certainly travel back along the beam, but in a disordered manner. Crossing of the electron trajectories will occur, and a region of confusion will creep along the beam from the ends at a velocity of at least $\eta BbI/I_0$ meters/sec., the mean value of $\dot{x}$ for an electron at the edge of the beam. We conjecture that if the beam travels so far in the z direction that $z \gg \dot{z}l/\dot{x}$, where l is the extent of the beam in the x direction, the shape becomes more and more oval, eventually becoming circular and increasing slowly in diameter, with completely disordered electron motion in the xy plane.

Conclusion

A simple idealized theory is able to give some guide to the properties of magnetically constrained strip and annular electron beams within certain well-defined limits. The main characteristics of the motion are qualitatively similar to those of cylindrical beams. If the current density drops to below half that required for Brillouin flow however the electron trajectories converge through a series of lines perpendicular to the short axis of the beam and the direction of electron flow. In practice such lines would become finite regions of increasing size and confusion. This phenomenon has no analog in cylindrical beams. The effect of the position and potential of the drift space walls is found to be important in the design of high perveance beams.

List of Symbols

The rationalized MKS system of units is used.

x_1, y_1 = co-ordinates of electron in the incident ($z=0$) plane.

Δz = distance traversed by outer electron of beam in crossing the y component of magnetic field (see Fig. 1).

$\dot{x}_1$ = x velocity of electron in $z=\Delta z$ plane.

d = y co-ordinate of plane of symmetry of magnetic field.

$\pm b$ = y co-ordinate of beam edge at incident plane.

$\pm g$ = y co-ordinate of conducting plane on which the electric field due to the beam terminates.

ϵ = permittivity of free space $= 8.85 \times 10^{-12}$ farads/meter.

η = charge to mass ratio of electron $= 1.76 \times 10^{11}$ coulombs/kgm.

B = magnetic induction in webers/square meter.

ρ = charge density in the electron beam, coulombs/cubic meter.

I = electron current in the z direction in amps/meter measured in the x direction.

$I_0 = 2\sqrt{2}\eta^{3/2}\epsilon V_0^{1/2}bB^2$ amps/meter $= 1.85 \times 10^6\ V_0^{1/2}bB^2$.

E = electric field in volts/meter.

V_0 = potential through which incident electrons have been accelerated from rest.

k = perveance/meter of incident beam.

$S = (1/2b) \times$ maximum or minimum beam thickness in y direction (whichever gives S different from unity).

V' = potential with respect to the beam edge of conducting planes on which the electric field from the beam terminates.

λ = distance in which beam contour repeats.

Transistor Shift Registers*

R. H. BAKER†, I. L. LEBOW†, AND R. E. McMAHON†

***Summary*—Three different types of high-speed transistor shift registers are discussed. The high speed (3 to 5 microseconds per shift pulse) is made possible by the use of nonsaturating bistable circuits. Two general shift registers are described making use of respectively one and two transistors per stage. The third register is of a specialized variety capable of shifting a single digit. In addition, an analysis is made of the triggering requirements of the register using a single transistor per stage.**

I. Introduction

IN GENERAL, shift registers are composed of a chain of interconnected bistable elements. When desired, the device may be made to perform such operations as sampling, coding, decoding, storing, etc.

This paper describes three different types of nonsaturating[1] bistable circuits, two of which may be used as the bistable elements necessary to construct shift registers. The third circuit offers a convenient method of building a sampling device such as a matrix switch. The shift-register logic is essentially the same as that used with vacuum tubes, that is, between each bistable stage is a diode "and" gate which controls the state of the succeeding stage.

In Section II, we present two of the basic circuits and describe the static bistable characteristics of each. Section III shows the shift-register logic used. Section IV gives an analysis of the single- and double-transistor bistable circuits. Section V discusses the specialized transistor shift register, and in Section VI we discuss the limitations of each as to speed, reliability, and so forth.

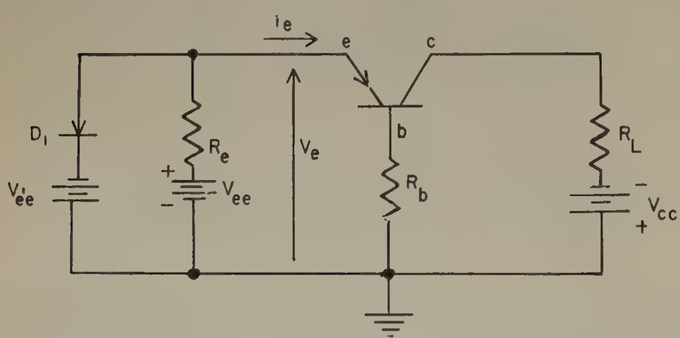

Fig. 1—Negative-resistance bistable circuit.

II. The Bistable Circuits[2]

The Negative-Resistance Bistable Circuit

In order to avoid the problems arising in saturated transistor circuits, the circuit of Fig. 1 was devised.[2]

The diode D_1, and resistor R_e, along with V_{ee} and V_{ee}', present the broken load line $R_e'R_e$ to the transistor emitter input characteristics. This is shown in Fig. 2. When the transistor is in the high conducting state (Point *b*, Fig. 2), the emitter voltage V_e is equal to V_{ef}. Point *b* is short-circuit unstable, therefore it is necessary that R_e be larger than the negative transistor input resistance $-R_N$. Moreover, the capacitance at the emitter must be kept small (see Section IV). The second point (*a*) occurs when the transistor is inactive. Hence we have a single-transistor bistable circuit with two nonsaturated stable states.

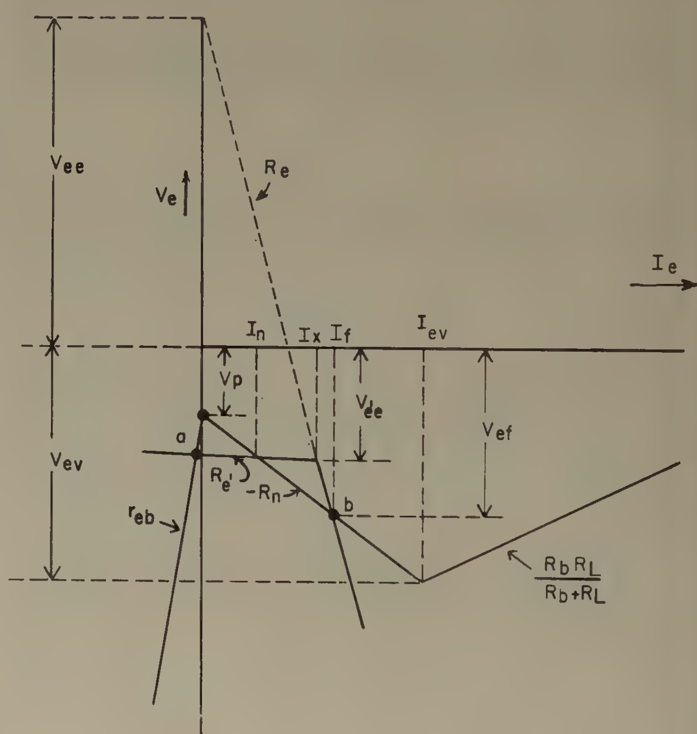

Fig. 2—Static emitter voltage-current characteristics.

The Two-Transistor Nonsaturating Flip-Flop[3]

The circuit shown in Fig. 3(a) was devised to avoid the effects of minority-carrier storage[4] that arise in saturated flip-flops. In order to explain the circuit characteristics, we break the circuit in a symmetrical way as in Fig. 3(b) and plot the input characteristics in Fig. 3(c). Points *a* and *b* are the only stable points since the positive input resistance of the "off" transistor in parallel with R_e exceeds in magnitude the negative input resistance of the "on" transistor.[2] Point *c*, where both transistors are on, is unstable since both transistors have negative input resistances. The current I is kept smaller than I_{ev} to avoid operation in the saturated region.

* Decimal classification: 621.375.2×R282.12. Original manuscript received by the IRE, October 27, 1953; revised manuscript received, February 17, 1954. Research in this document was supported jointly by the Army, Navy, and Air Force under contract with Mass. Inst. of Tech.

† Lincoln Lab., Mass. Inst. of Tech., Cambridge, Mass.

[1] A bistable circuit that has active points in the negative-resistance region.

[2] I. L. Lebow, R. H. Baker, R. E. McMahon, "The Transient Response of Transistor Switching Circuits," Technical Report No. 27, Lincoln Laboratory, M.I.T., July, 1953.

[3] A. W. Carlson, "A Transistor Flip-Flop with Two Stable Nonsaturating States," AFCRC Report, December, 1952.

[4] R. A. Bradbury, "Hole Storage or Turn Off Time," AFCRC, December, 1952.

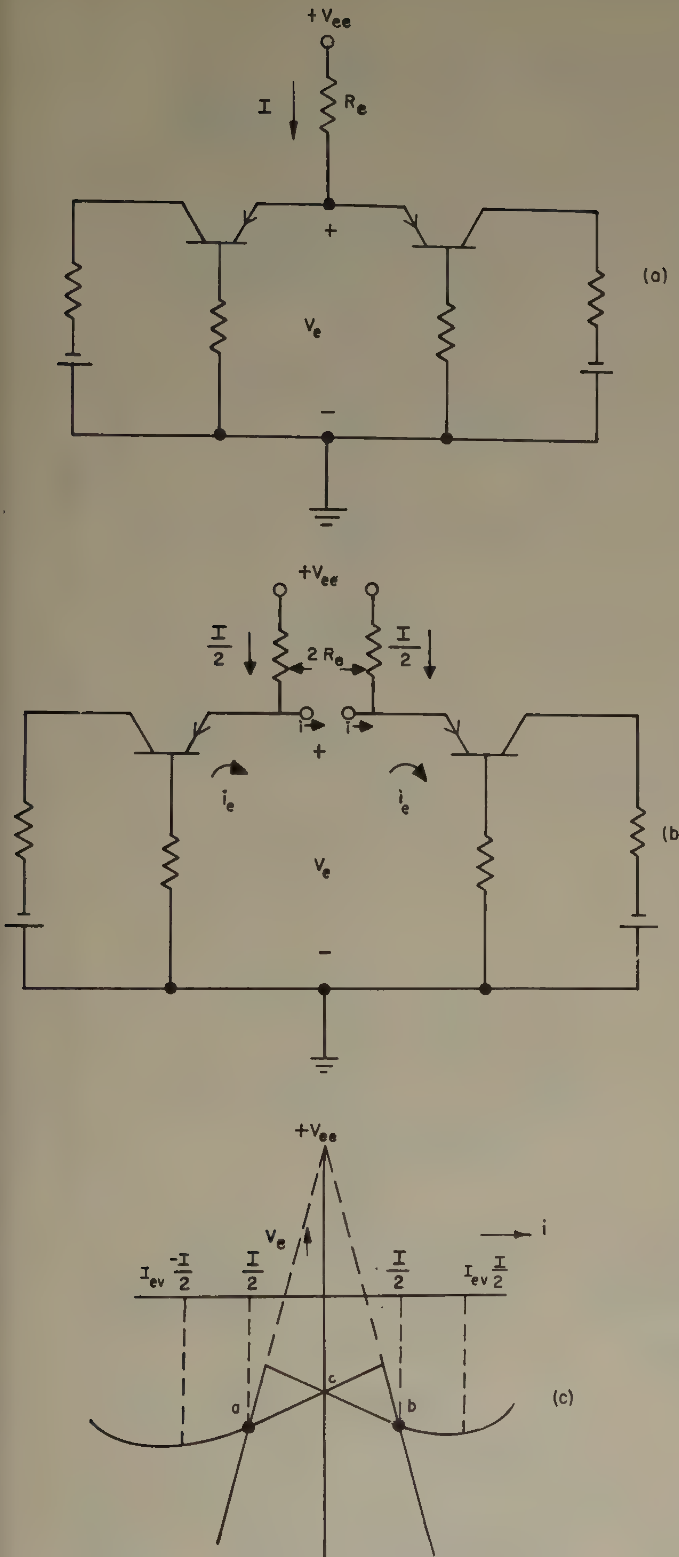

Fig. 3—Nonsaturating flip-flop.

III. The Transistor Shift Register

Single Transistor Per Stage

A shift register may be formed from the bistable element discussed in Section II by inserting diode gates between each stage. This is shown in Fig. 4. The first stage is "set" (switched from Point *a* to Point *b*, Fig. 2) by clock pulses that are controlled by the input binary code. When Stage 1 is set, the diode gate allows a clock pulse to set the second stage. Stage 2 "resets" Stage 1 and allows the next clock pulse to set the third stage, etc. In this manner, the "sense" of the first stage transfers down the chain of bistable elements.

For a detailed three-stage schematic of the shift register, see Fig. 5, Page 1154. Fig. 6, Page 1155, is an analysis of the wave-forms at various points of the circuit.

The Flip-Flop Shift Register

The flip-flop shift register is formed from the bistable elements in the same way as with vacuum-tube shift registers, that is, the inputs to a particular stage are the clock pulses that have been gated from the outputs of the previous stage; this is shown in Fig. 7 on page 1156.

IV. Analysis and Design of the One- and Two-Transistor Bistable Circuits

There are several factors that complicate the design of the transistor bistable circuits necessary to construct high-speed shift registers. First, the variations in transistor parameters from unit to unit necessitate that the basic design of the bistable circuit be stable over quite large percentage changes in operating points. Second, since most presently available transistors (Type BTL 1689) have relatively low-frequency cutoffs (less than 5 Mc), the trigger pulse must be controlled in both amplitude and width.[2,5] Third, since the input impedance of a point-contact transistor circuit is, in general, lower than output impedance, cascading stages becomes difficult.

From Figs. 5 and 6 we see that the "reset" capacitor (Point *h*) must recover through the series combination of resistances R_2 and R_s. Thus, to increase the maximum operating speed, the triggering capacitors must be made as small as possible. To determine the minimum value capacitors that will suffice to reliably switch the nonsaturating bistable circuits, a transient analysis must be made. This analysis is based upon the methods described in detail in the literature. The model is that of the large signal equivalent circuit of Adler[6] with the dynamic interpretation shown in Fig. 8 on page 1157.

Stability Conditions

In all cases we are considering circuits whose "on" stable point is in the active region of the transistor characteristic. The transistor, looking from emitter to ground, displays a negative input impedance and is short-circuit unstable. For stability the external emitter resistance must exceed in magnitude the negative input resistance. In addition, when the external emitter resistance is large, the capacitance between emitter and ground C_e, must satisfy the relation

$$C_e < \frac{\tau}{R_n} = \frac{1}{2\pi f_{co} R_N} \tag{1}$$

[5] R. H. Baker, "Transistor Shift Register," M. S. Thesis, M.I.T., June, 1953 (E.E. Dept.).

[6] R. B. Adler, "A Large Signal Equivalent Circuit for Transistor Static Characteristics," M.I.T., R. L. E. Transistor Group Report T-2, August, 1951.

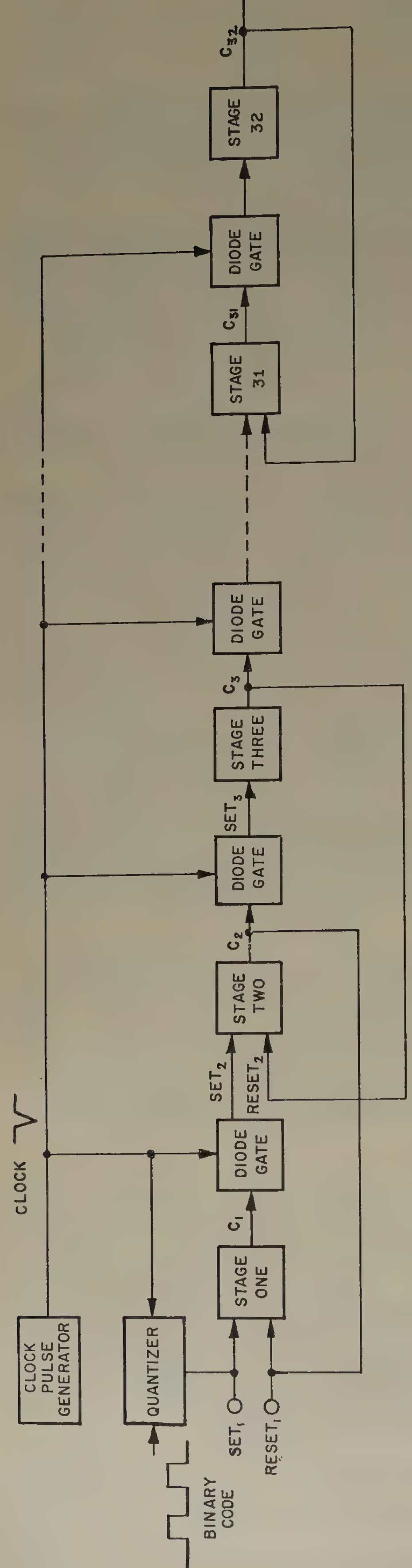

Fig. 4—Block diagram of single-transistor-per-stage shift register.

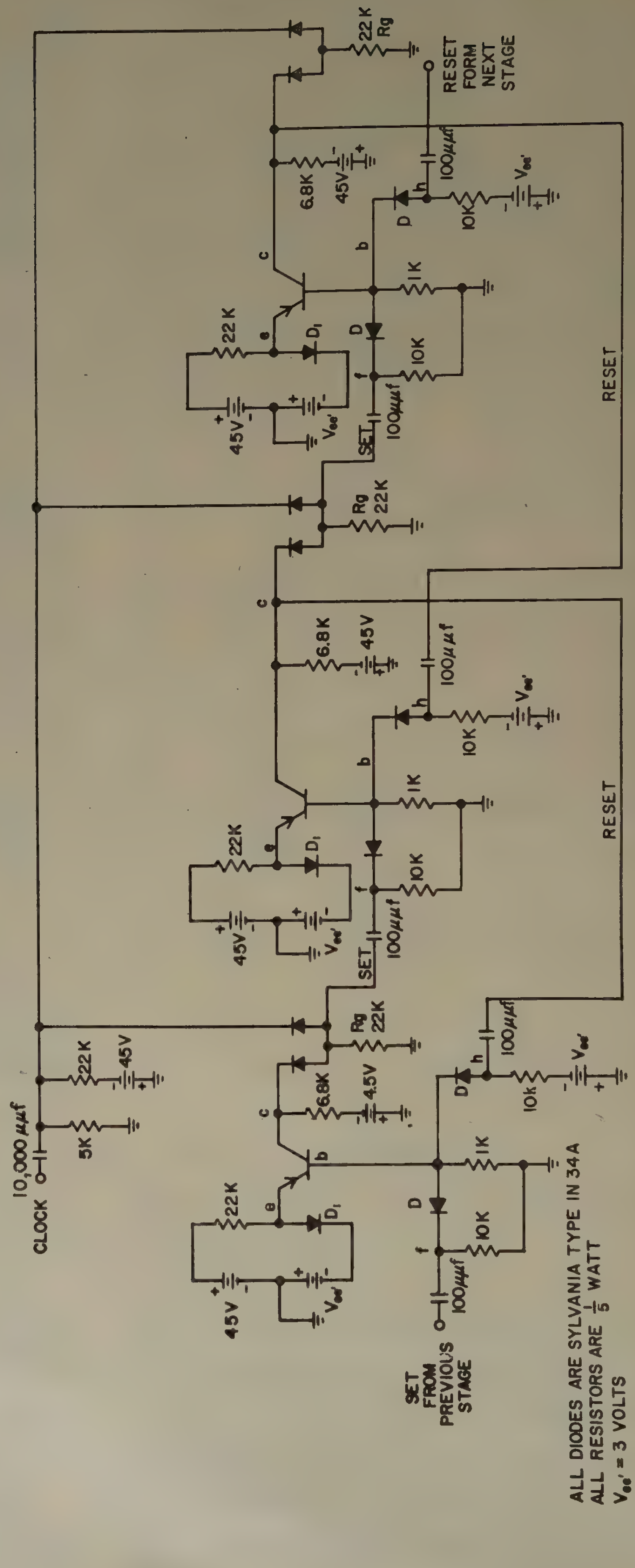

Fig. 5—Circuit diagram of single-transistor-per-stage shift register.

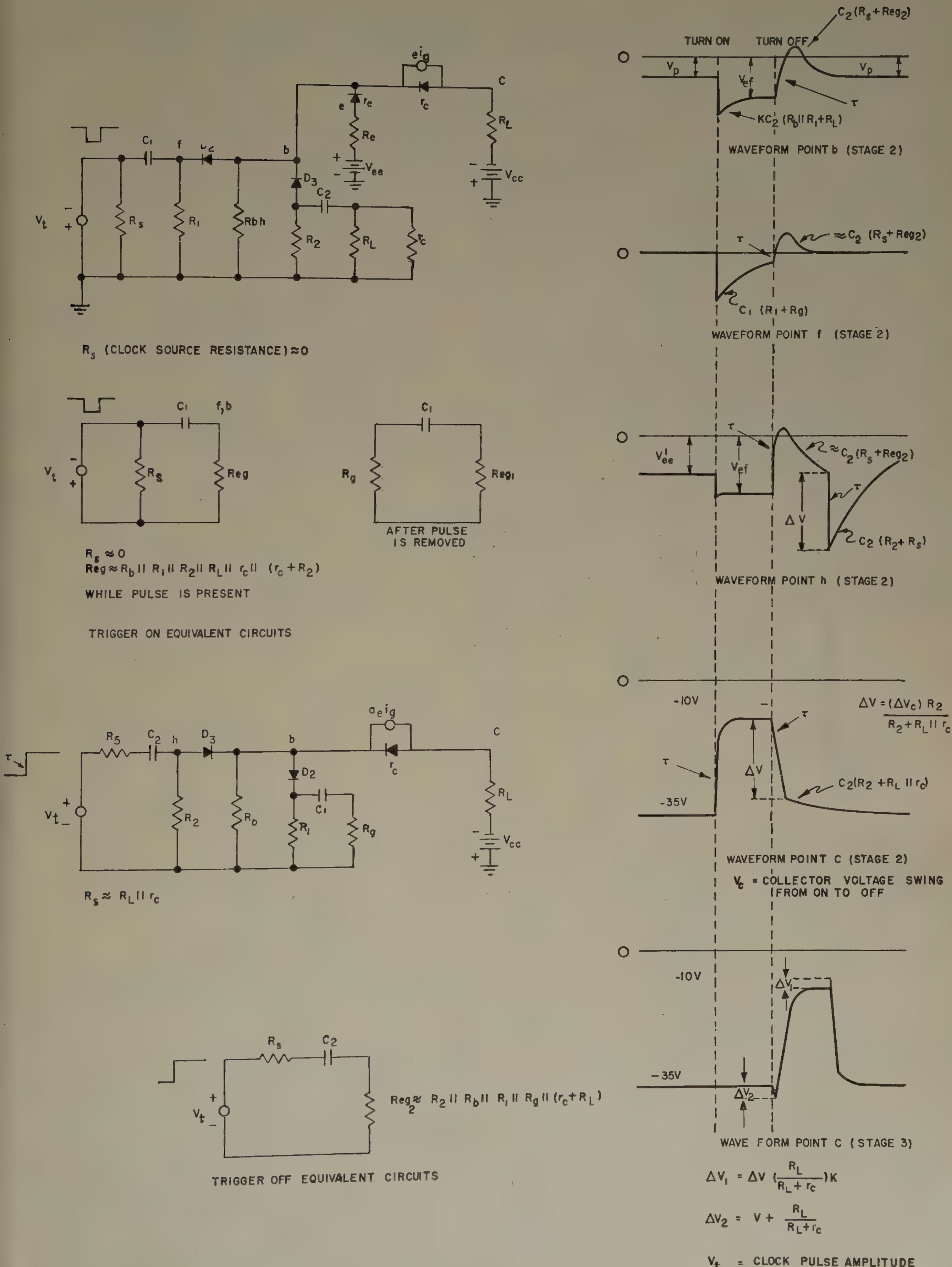

Fig. 6—Waveform analysis of single-transistor-per-stage shift register.

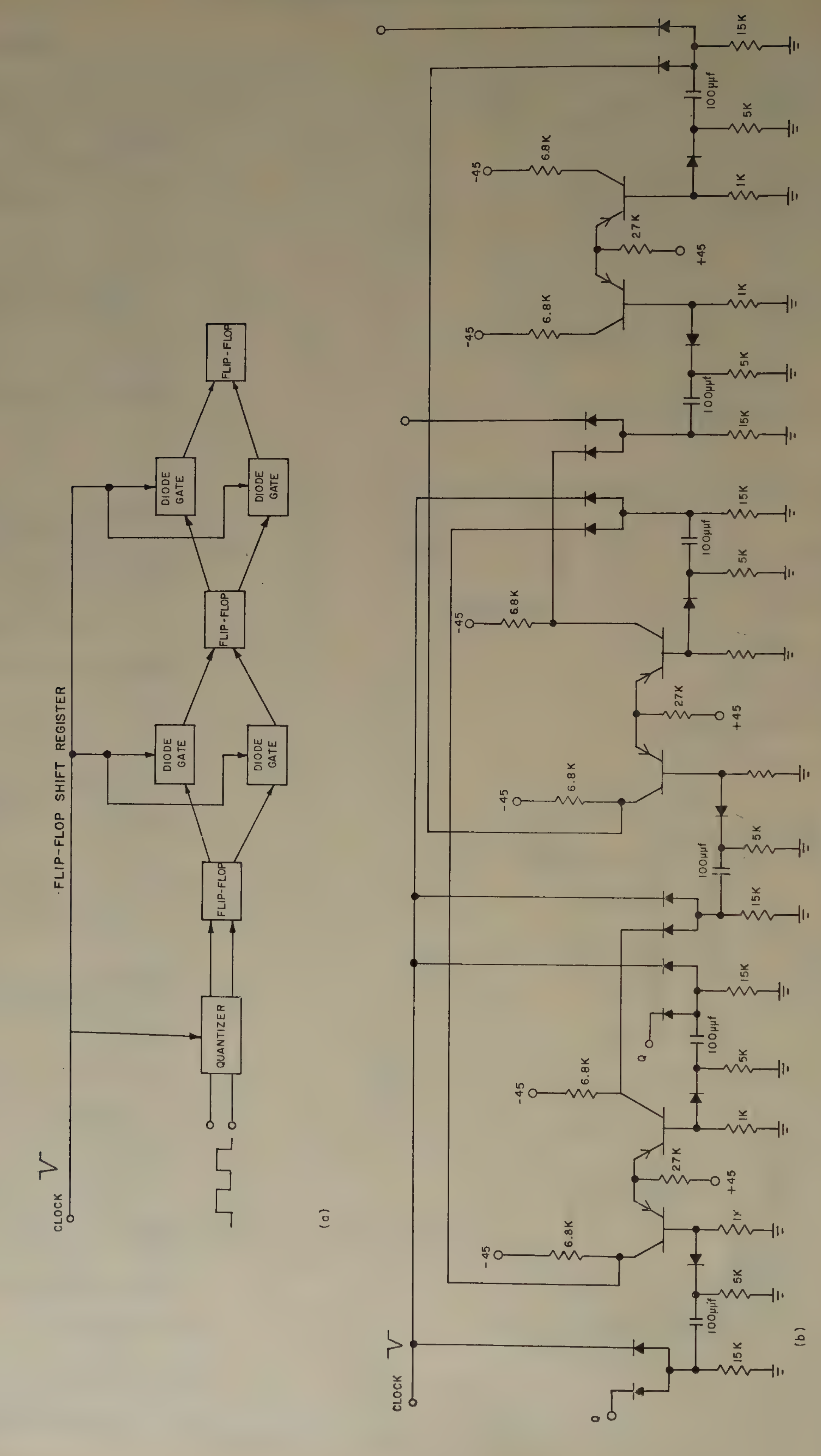

Fig. 7—Flip-flop shift register. (a) Block diagram; (b) Circuit diagram.

where f_{co} is the transistor cutoff frequency. The above relation is obtained[2,7] by demanding that the transient currents in the active region be converging exponentials.

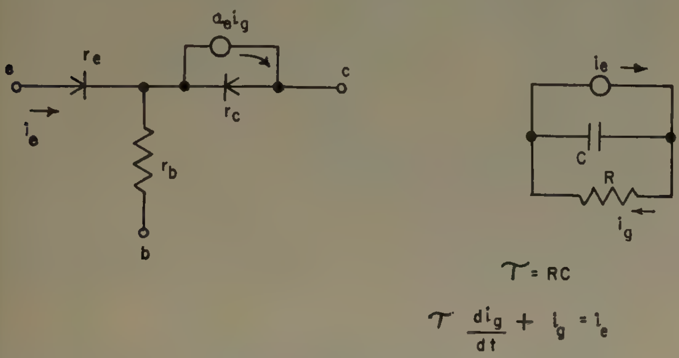

Fig. 8—Large-signal equivalent circuit for transistor dynamic characteristics.

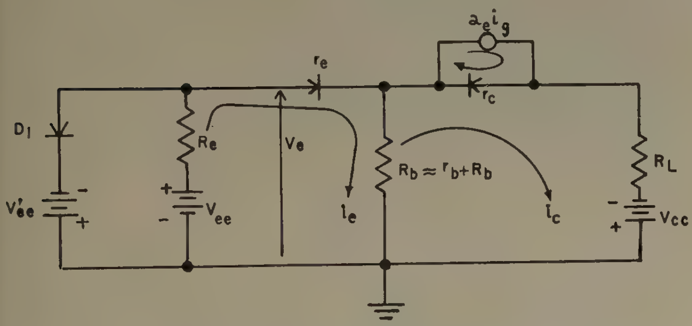

Fig. 9—Negative-resistance bistable circuit with transistor large-signal dynamic equivalent circuit.

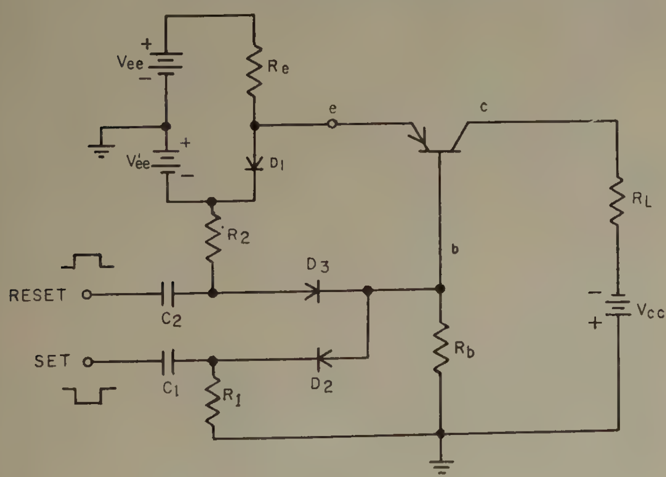

Fig. 10—Negative-resistance bistable circuit.

Triggering Requirements of the Negative-Resistance Bistable Circuit

Turn On: Consider the circuit shown in Fig. 10. In order to be sure that the negative-resistance bistable circuit will switch (change states) reliably, it is necessary that the trigger pulse exceed a given amplitude for a given period of time.[2,5] Moreover, the amplitude-width requirement is different depending upon the state of the circuit[2,5] and the frequency response of the transistor. Consider only the "set" (turn on) function of the circuit shown in Fig. 10.

[7] B. G. Farley, "Dynamics of Transistor Negative Resistance Circuits," Proc. I.R.E., vol. 40, pp. 1497–1507; November, 1952.

Solving the loop equations, the following expression may be obtained for i_g:

$$\frac{d^2 i_g}{dt^2} + \frac{di_g}{dt}\left(\frac{1}{\tau} + \frac{R_p + R_e}{C_1 R_e R_p}\right) + i_g\left(\frac{R_e - R_N}{C_1 R_e R_p \tau}\right) - \frac{V_{cc} + V_p}{C_1 R_e R_p \tau}, \qquad (2)$$

where R_p is the input resistance of the transistor with $\alpha = 0$, and V_p is the magnitude of the peak point voltage. The solution of (2) is the sum of two exponentials, the frequencies of which are given by

$$\omega^2 + \omega\left(\frac{1}{\tau} + \frac{R_e + R_p}{C_1 R_e R_p}\right) + \frac{R_e' - R_N}{C_1 R_e R_p \tau} = 0. \qquad (3)$$

The applied set pulse opens D_1. R_e is greater than R_N. Hence, from (3) both ω's are negative. The initial shape of i_e then, is a positive step, the amplitude of which depends upon the applied trigger voltage, V_t, followed by a converging decay. Since i_g tries to follow i_e, the eventual circuit state is dependent upon the relative magnitudes of i_e and i_g as time progresses. For when i_e decays to I_x (Fig. 2), diode D_1 closes and R_e becomes R_e'. Since R_e' is less than R_N, the third term in (3) is negative and the circuit is unstable. In order for the circuit still to switch on, even though i_e becomes less than the value I_x, the rate of change of i_e must, at some time after t_x, be equal to zero (t_x is defined as the time at which i_e is equal to the value I_x).

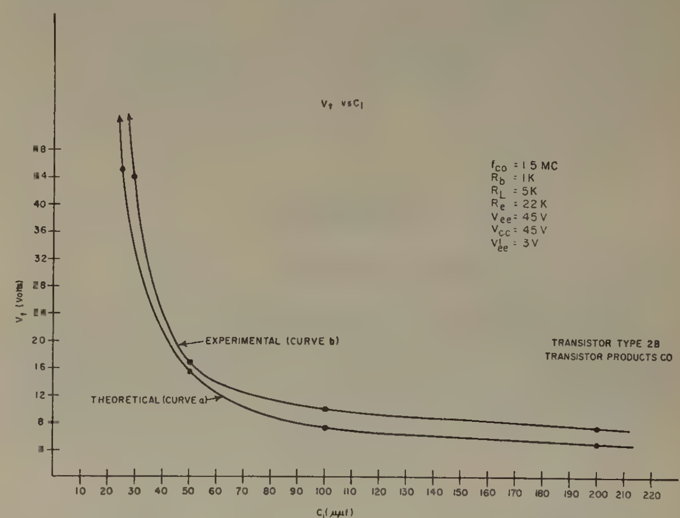

Fig. 11—Turn-on characteristic of negative-resistance bistable circuit.

The approximate criterion for triggering on is[2,5]

$$i_g(t_x) > I_N \text{ Turn on} \qquad (4a)$$

$$i_g(t_x) < I_N \text{ Turn back off.} \qquad (4b)$$

The values of C_1 as a function of the trigger voltage V_t that will fulfill the condition (4a) appear in Fig. 11. The results are found to be in good agreement with experimental data.

Turn-Off: The approach that may be used to determine the "turn-off" trigger requirement is the same

as for the turn-on case; using the appropriate equivalent circuit, we solve the pertinent loop equations, define the turn-off criterion, and solve for the minimum trigger voltage that will suffice to switch the circuit.

Since the circuit is turned off by applying a positive pulse to the base, the transistor emitter is instantaneously switched off and C_2 starts charging (Fig. 10). When the base voltage decays below V_{ee}', i_e starts increasing. The same arguments as in the previous section show that, in order for the circuit to switch off, i_e must go to zero. As in the case of turning on, the triggering diode D_3 will open when the rate of change of i_e is equal to zero. Hence at this time t_Δ,[2,5]

$$i_g(t_\Delta) > I_N \text{ Turn back on} \tag{5a}$$

$$i_g(t_\Delta) < I_N \text{ Turn off.} \tag{5b}$$

Fig. 12 shows a comparison between the theoretical and experimental data when V_t is plotted as a function of C_2.

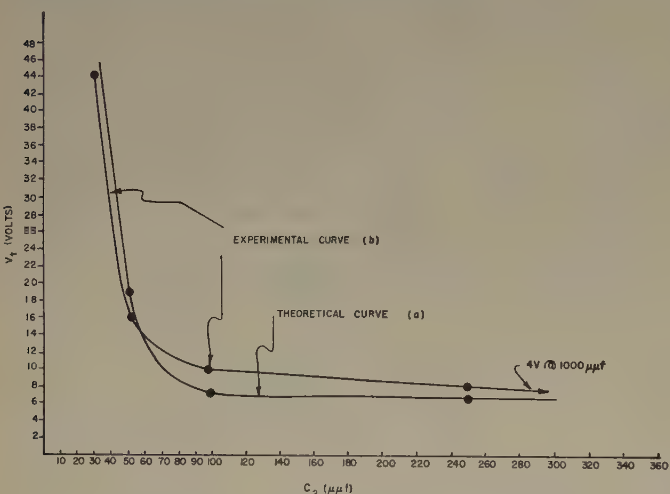

Fig. 12—Turn-off characteristics of negative-resistance bistable circuit.

The Two-Transistor Nonsaturating Flip-Flop[4]

We may replace the transistors in the circuit of Fig. 3(a) by the equivalent circuit shown in Fig. 8. When this is done, the circuit of Fig. 13 is obtained, where the battery V_t implies that we are triggering on transistor B by applying a negative pulse to the base.

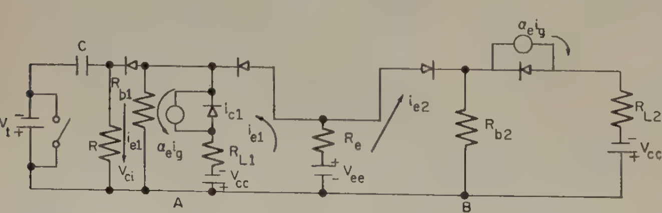

Fig. 13—Dynamic equivalent circuit of two-transistor flip-flop.

The circuit of Fig. 13 may be analyzed by the same techniques described above. The solutions for the various current equations will be of the same form, although slightly more complex due to the circuit configuration. As above, the triggering requirements have a definite minimum amplitude-width relationship. This analysis is made elsewhere,[8] therefore it will not be repeated. However, an important point should be noted: the two-transistor flip-flop will trigger on narrower pulses than will the negative-resistance bistable circuit. This is because, although we are triggering at the base of transistor A (Fig. 13), and hence, the same conditions apply as when triggering the single transistor, transistor B is being triggered at the emitter. Since the circuit is short-circuit unstable at the emitter, the currents in transistor B decay (or increase, as the case may be) with positive exponentials. This causes faster switching of transistor B, which in turn will cause the circuit to switch faster.

A Specialized Transistor Shift Register[7]

A logical extension of the circuit shown in Fig. 3(a) is the multistage circuit of Fig. 14 where several transistor stages have their emitters tied to a common load resistor. As in the two-transistor case, only one transistor can conduct at a time, since any state with more than one transistor in the negative-resistance region is unstable.

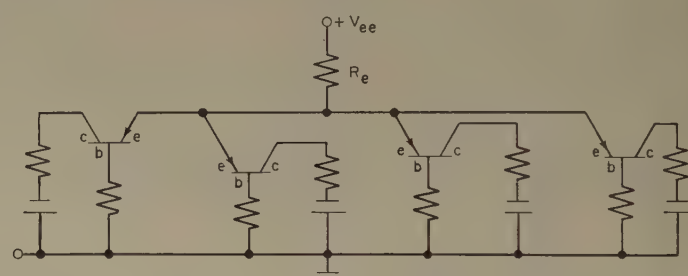

Fig. 14—Four-transistor circuit with four stable states.

A specialized shift register may be based upon this idea, as shown in Fig. 15. Suppose a negative pulse is applied to the base of the first stage. This stage will then be conducting and the other nonconducting. The sense of this first may be then propagated down the register by clock pulses gated to stages sequentially.

The number of stages that may be connected together is limited by two factors. First, the total accumulated capacitance between emitter and ground must be kept small enough to render each transistor stable (1). Second, the resistance between emitter and ground, which is reduced by the addition of more transistors, must be kept large enough to render each transistor's "on" operating point in the negative-resistance region.

As many as ten transistors may be connected in this manner with little or no difficulty. We may construct a long register, using this method, as shown in Fig. 16.

Comparison and Evaluation

An over-all analysis of the shift-register circuits reveals the following facts: first, a point-contact transistor switching circuit requires a definite, predictable, minimum pulse-height width relationship that is a function

[8] R. E. McMahon, I. L. Lebow, R. H. Baker, "A Two Transistor Shift Register," Lincoln Laboratory, M.I.T., M24-20, May, 1953.

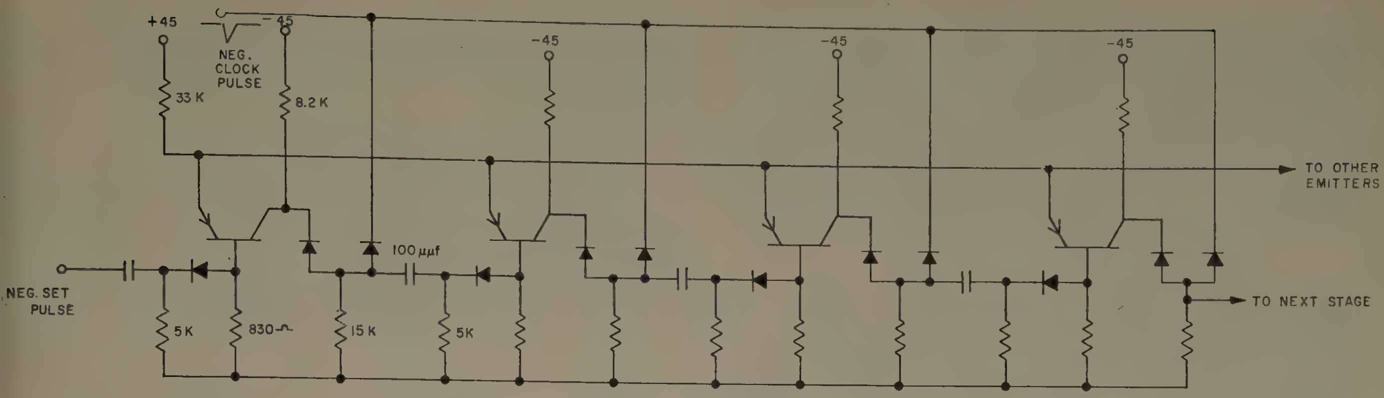

Fig. 15—Circuit diagram of specialized shift register.

of the transistor frequency response and the circuit configuration; secondly, the maximum speed of operation of a system constructed from these circuits is limited, not by the switching time of the circuits, but by the recovery time of the associated triggering capacitors. In view of this, when designing high-speed systems, it is advantageous to use circuits with (a) high input impedance, and (b) as low recovery resistance as possible for the triggering-capacitor circuits.

The above results appear to exclude the possibility of using saturated bistable circuits with a high degree of reliability due to the turn-off triggering time.[2]

Exhaustive tests have been made on each of the shift registers described above. The two-transistor flip-flop works reliably with 80 per cent of all BTL 1698 transistors at rates up to 5 μsec per shift pulse. The shift register constructed from the negative-resistance bistable circuit will operate up to 6 μsec per stage.

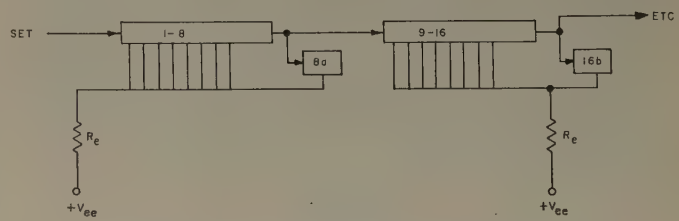

Fig. 16—Block diagram of specialized shift register showing arrangement for designing long shift registers.

However, at these fast rates, the coupling capacitors must necessarily be small, and this shift register therefore is less reliable than the flip-flop shift register. At lower rates, the coupling capacitor may be larger with correspondingly greater reliability. The specialized shift registers offer good reliability up to 3 μsec per shift pulse when a maximum of 9 stages is connected to the common bias resistor, R_e. Longer registers should be connected as shown in Fig. 16.

Comb Filters for Pulsed Radar Use*

S. F. GEORGE†, SENIOR MEMBER, IRE AND A. S. ZAMANAKOS†

Summary—**This paper presents a discussion of the concepts involved in the use of "comb" filters for signal-to-noise improvement in pulsed radar systems. A "comb" filter is one whose frequency spectrum consists of a number of equi-spaced elements resembling the tines of a comb. Consideration is given to the characteristics of the frequency spectra arising from sequences of pulses varying in number from one to infinity. These results indicate that the use of comb filters may be feasible in most cases. Hence, a study is made of the use of three different types of filters to determine their effect on the signal-to-noise ratio. These filters range from the simplest type of uniform filters to the North type of matched filter. The relative merits of these filters are determined in terms of improvement in signal-to-noise ratio.**

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† Naval Research Laboratory, Washington, D. C.

Introduction

A PAPER which appeared in the literature[1] on the detection of periodic signals in noise prompted a spirited discussion[2] which centered about the characteristics of the frequency spectra of pulsed signals and the feasibility of improving the signal-to-noise ratio by the use of comb filters. The question of the frequency spectrum of the received signal is of fundamental im-

[1] Y. W. Lee, T. P. Cheatham, Jr., and J. B. Wiesner, "Application of correlation analysis to the detection of periodic signals in noise," Proc. I.R.E., vol. 38, pp. 1165–1171; October, 1950.

[2] Discussion on "Application of correlation analysis to the detection of periodic signals in noise," by Lee, Cheatham, Jr., and Wiesner, with N. Marchand, M. Leifer, and H. R. Holloway, Proc. I.R.E., vol. 39, pp. 1094–1095; September, 1951.

portance in radar systems and has a direct bearing on the effectiveness of using comb-filter techniques to extract the signal from additive white noise. It is the purpose of this paper to show how the frequency spectra vary as the number of pulses received ranges from a very few to an appreciable number and then to clarify the concepts involved in the use of various comb filters.

Spectra of Pulse Trains

The frequency spectrum of a single pulse is well known and has a continuous distribution of sin x/x shape. It is also well known that the spectrum of an infinite sequence of equi-spaced pulses, of uniform height and width, consists of discrete lines. This spectrum as illustrated in Fig. 1 has a sin x/x envelope. The discrete lines, often called spectral or pulse-repetition-frequency lines, are located at those points along the angular frequency axis which are integral multiples of $2\pi/\tau$; where τ is the pulse-repetition period.

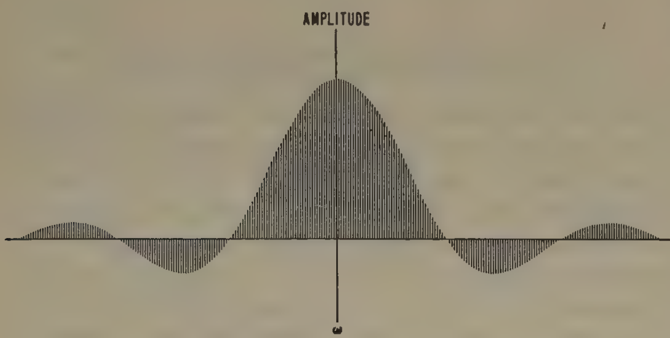

Fig. 1—Discrete frequency spectrum of an infinite pulse train.

The general frequency spectrum for a finite number of pulses, the case normally encountered in search radar operation, will next be considered. Let us assume a pulse train of m pulses of height A, pulse width δ and pulse period τ. There will be no loss of generality if we select an odd number of pulses, say $m=2k+1$, in order to simplify the mathematics. The frequency spectrum for this finite pulse sequence is given by the Fourier transform,

$$F(\omega) = \frac{1}{2\pi}\int_{-\infty}^{\infty} f(t)e^{-j\omega t}dt, \tag{1}$$

where $f(t)$ defines the pulses in real time. Evaluating this as an even function we have

$$F(\omega) = \frac{1}{\pi}\int_{0}^{\infty} f(t)\cos\omega t dt$$

$$= \frac{A\delta}{2\pi}\,\frac{\sin\frac{\omega\delta}{2}}{\frac{\omega\delta}{2}}\,\frac{\sin\frac{m\omega\tau}{2}}{\sin\frac{\omega\tau}{2}}. \tag{2}$$

In order to show how the continuous frequency distributions change as the number of received pulses varies, let $F(\omega)$ be normalized, and let $\delta=1$ and $\tau=3$, giving

$$F_n(\omega) = \frac{\sin\frac{\omega}{2}}{\frac{\omega}{2}}\,\frac{\sin\frac{3m\omega}{2}}{\sin\frac{3\omega}{2}}. \tag{3}$$

Fig. 2 shows that for as few as 19 pulses there is a tendency for the energy to be concentrated about those frequencies where prf lines would be located if there were an infinite pulse train. When m is increased to 999 as in Fig. 3 this tendency becomes so pronounced as to present essentially the same spectrum as for an infinite number of pulses. Note that the envelope of all these plots has the same sin x/x shape with the maximum amplitude increasing with the number of pulses received. It will now be demonstrated mathematically how a comb type of filter might be employed to improve the signal-to-noise ratio.

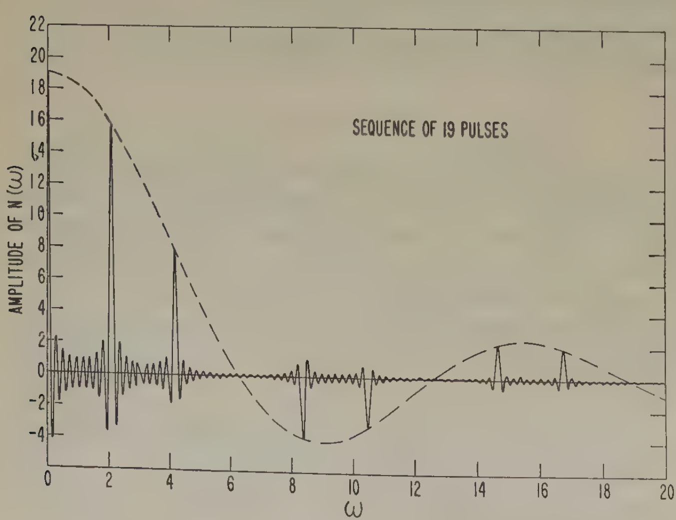

Fig. 2—Frequency response of a sequence of 19 pulses.

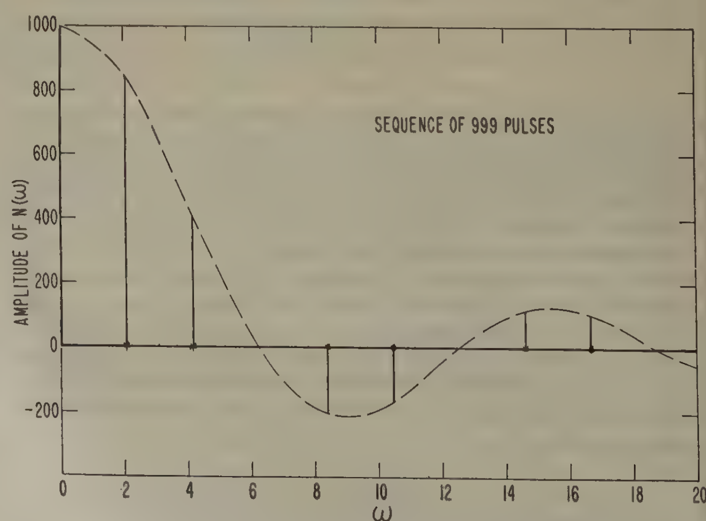

Fig. 3—Frequency response of a sequence of 999 pulses.

Basic Formula for Comb Filters

In general if we have a function, $f(t)$, which is known to be zero outside a given interval of time and which may assume any finite form within this specified interval, we may represent this function in two ways: 1. by the Fourier transform, if the function belongs to L^1 [i.e., $\int |f(t)| dt < \infty$ over $(-\infty, \infty)$] and 2. by a Fourier series which is valid only in the prescribed interval. The first of these was used to develop (2) since $f(t)$ was in L^1. The second will be used to develop the following theorem:

Theorem: Let $f(t)$ be in L^1 and be zero except in $-T \leqq t \leqq T$ and let

$$F(\omega) = \frac{1}{2\pi}\int_{-\infty}^{\infty} f(t)e^{-j\omega t}dt.$$

Then

$$F(\omega) = \sum_{n=-\infty}^{\infty} \chi_n \frac{\sin T\left(\omega - \frac{n\pi}{T}\right)}{T\left(\omega - \frac{n\pi}{T}\right)}, \tag{4}$$

where

$$\chi_n = F\left(\frac{n\pi}{T}\right).$$

The proof of this theorem follows.

For a function, $f(t)$, as defined, the Fourier transform is

$$F(\omega) = \frac{1}{2\pi}\int_{-T}^{T} f(t)e^{-j\omega t}dt. \tag{5}$$

Since $f(t)$ can be specified to be a periodic function of period $2T$ we may develop it into a Fourier series:

$$f(t) = \sum_{n=-\infty}^{\infty} C_n e^{j(n\pi t/T)} \tag{6}$$

where

$$C_n = \frac{1}{2T}\int_{-T}^{T} f(t)e^{-j(n\pi t/T)}dt. \tag{7}$$

Using (6) in (5) and integrating:

$$F(\omega) = \sum_{n=-\infty}^{\infty} C_n \frac{T}{\pi} \frac{\sin T\left(\omega - \frac{n\pi}{T}\right)}{T\left(\omega - \frac{n\pi}{T}\right)}. \tag{8}$$

From (5) and (7) we see that

$$C_n = \frac{\pi}{T} F\left(\frac{n\pi}{T}\right)$$

and hence (8) becomes

$$F(\omega) = \sum_{n=-\infty}^{\infty} F\left(\frac{n\pi}{T}\right) \frac{\sin T\left(\omega - \frac{n\pi}{T}\right)}{T\left(\omega - \frac{n\pi}{T}\right)}. \tag{9}$$

If we let $T = m\tau/2$ in (9), then

$$F(\omega) = \sum_{K=-\infty}^{\infty} F\left(\frac{2\pi K}{\tau}\right) \frac{\sin \frac{m\tau}{2}\left(\omega - \frac{2\pi K}{\tau}\right)}{\frac{m\tau}{2}\left(\omega - \frac{2\pi K}{\tau}\right)}, \tag{10}$$

where $K = n/m$ will be found to take on only integral values. This change of index is valid since at all points where K is not an integer the coefficient $F(2\pi K/\tau)$ will be zero.

A study of (10) shows frequency spectrum of a finite sequence of pulses consists of a sum of $\sin x/x$ curves whose height at the center of each curve is equal to the height of the corresponding spectral line in the spectrum of an infinite sequence of pulses whose amplitudes are multiplied by m. Because the form of (10) so well defines the spectrum and indicates that comb filters are feasible, it is called the basic formula for comb filters.

Discussion of Three Types of Comb Filters

Most previous discussions on comb filters have been concerned with the case of an infinite train of pulses giving rise to a discrete line spectrum. The basic idea is to place as narrow band-pass filters as feasible about these spectral lines, passing the signal energy and removing as much noise as possible. However, Fig. 2 showed that in the practical case for a finite sequence of pulses spectrum is continuous and has a definite frequency spread about those points where prf lines would occur for the case of an infinite pulse train. This brings up the question of proper bandwidth of individual comb filters. One approach is to consider the energy present in a specified frequency interval as compared to total energy over the entire spectrum. Let us take this specified interval to include the main hump of energy distribution; i.e., for $-(2\pi/\delta) \leqq \omega \leqq (2\pi/\delta)$ where δ is pulse width.

The total energy in the spectrum of a single pulse is

$$E_T = 2\pi \int_{-\infty}^{\infty} |F(\omega)|^2 d\omega$$

$$= 2\pi\left(\frac{A\delta}{2\pi}\right)^2 \int_{-\infty}^{\infty} \left[\frac{\sin\frac{\omega\delta}{2}}{\frac{\omega\delta}{2}}\right]^2 d\omega = A^2\delta, \tag{11}$$

whereas the energy in the main hump is

$$E' = 2\pi\left(\frac{A\delta}{2\pi}\right)^2 \int_{-2\pi/\delta}^{2\pi/\delta} \left[\frac{\sin\frac{\omega\delta}{2}}{\frac{\omega\delta}{2}}\right]^2 d\omega = 0.90A^2\delta. \tag{12}$$

In other words, 90 per cent of the signal energy is concentrated about the center of the sin x/x pattern. The choice of this particular interval, then, for the width of our comb filters will provide efficient use of the available energy as well as simplifying the mathematics. It should be pointed out that this is merely one choice out of a number, and it may not be the best for optimizing the signal-to-noise ratio. The problem of interference between the side lobes of one sin x/x pattern and the main lobe of another will not be considered.

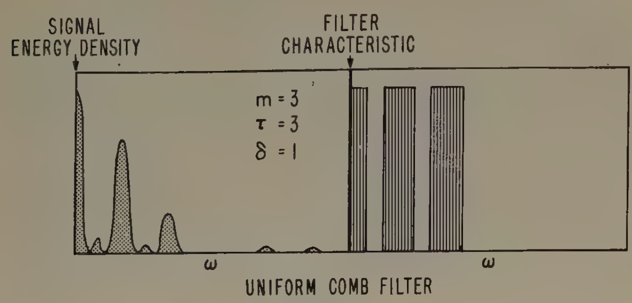

Fig. 4—Signal-energy density for a three-pulse sequence and uniform comb-filter characteristic.

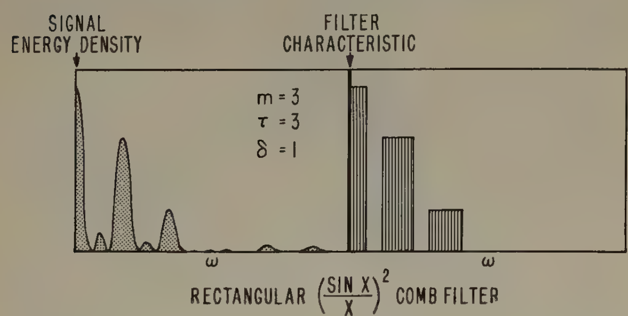

Fig. 5—Signal-energy density for a three-pulse sequence and rectangular(sin $x/x)^2$ comb-filter characteristic.

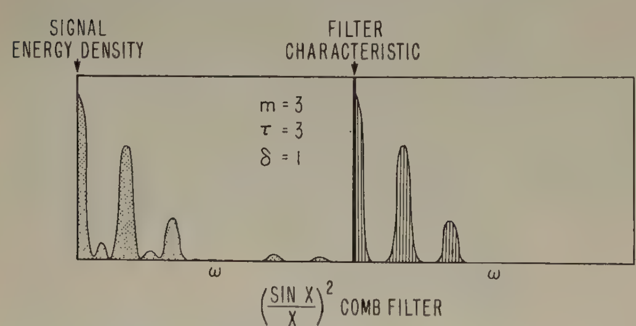

Fig. 6—Signal-energy density for a three-pulse sequence and North-type (sin $x/x)^2$ comb-filter characteristic.

Three types of comb filters will be analyzed: 1. the uniform comb filter of Fig. 4, 2. the rectangular (sin $x/x)^2$ comb filter (Fig. 5), and 3. the North-type (sin $x/x)^2$ comb filter (Fig. 6). The uniform comb filter was selected because it is the one usually thought of in connection with comb filtering. The so-called "rectangular (sin $x/x)^2$" comb filter was an attempt to improve the signal-to-noise ratio. The North-type[3] (sin $x/x)^2$ filter is the result of applying the North matched-filter technique in the frequency interval considered. The transfer characteristic derived by North to give the maximum signal-to-noise ratio is one that, except for a necessary time-shift factor, looks like the complex conjugate of the spectrum of the input signal under consideration.[4] In the case at hand the filter was shaped to the same form as the input signal.

In order to establish the degree of improvement obtained by the use of comb filters additive white noise is introduced into the picture. For band-limited white noise of total mean energy N_o in the interval $-(2\pi/\delta) \leq \omega \leq (2\pi/\delta)$ and of mean energy per unit bandwidth N_u, we have the following relation:

$$N_o = \frac{4\pi}{\delta} N_u. \tag{13}$$

A noise energy concept is used rather than noise power since the noise is gated to exist only for the duration of the pulse train with which we are concerned. To simplify the terminology say that the center point of each comb filter is located at a prf line; that is, at that point where the prf line would fall were there an infinite sequence of pulses. To get a better idea of how each element in the comb filter acts, Fig. 7 shows the signal- and noise-energy-density spectra about a single prf line and Fig. 8 shows two filter frequency-response curves. Note in Fig. 7 that the signal energy follows the (sin $x/x)^2$ pattern whereas the noise energy is flat. Also note that each filter bandwidth is $4\pi/m\tau$, a function of the number of pulses received in a sequence.

A. Type I: Uniform Comb Filter

Let the input signal-to-noise energy ratio be denoted by r_i. Then for a total input signal energy of E we have

$$r_i = 10 \log \frac{E}{N_o} \tag{14}$$

expressed in decibels. Now a set of uniform comb filters will leave 90 per cent of the energy about each prf line and hence the new filtered signal energy is 0.90 aE, where "a" represents the filter gain. To obtain the noise energy refer to Fig. 7. Here the bandwidth for each filter is $4\pi/m\tau$ and there are $[2(\tau/\delta)+1]$ filters of gain "a" in the interval under consideration, hence the noise output from the comb filters is

$$N_{CF} = a \frac{4\pi}{m\tau}\left(2 \frac{\tau}{\delta} + 1\right) N_u, \tag{15}$$

which from (13) becomes

$$N_{CF} = a\left(\frac{2}{m} + \frac{\delta}{m\tau}\right) N_o. \tag{16}$$

In most practical cases $(\delta/\tau) \ll 2$ and hence $N_{CF} \cong a(2N_o/m)$. The output signal-to-noise energy ratio for

[3] The expression "North-type" filter is used here since this filter covers only a prescribed portion of the frequency spectrum, whereas the usual North filter does not have this limitation.

[4] D. O. North, "Analysis of Factors which Determine Signal-Noise Discrimination in Pulsed Carrier Systems," unpublished report PTR-6C, RCA, Princeton, N. J.; June, 1943.

a Type I comb filter will be

$$r_{0I} = 10 \log \frac{0.90E}{2\,\dfrac{N_o}{m}} = r_i + 10 \log 0.45m. \qquad (17)$$

This indicates that the signal-to-noise energy ratio may be improved by at least 10 log 0.45 m db by virtue of comb filtering.

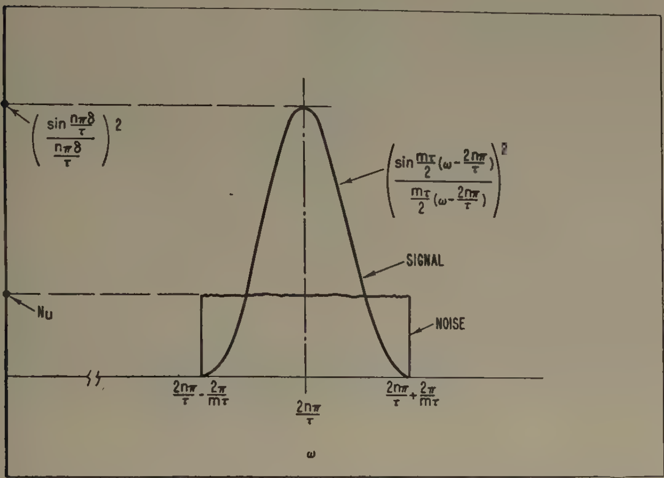

Fig. 7—Typical section of signal- and noise-frequency spectra.

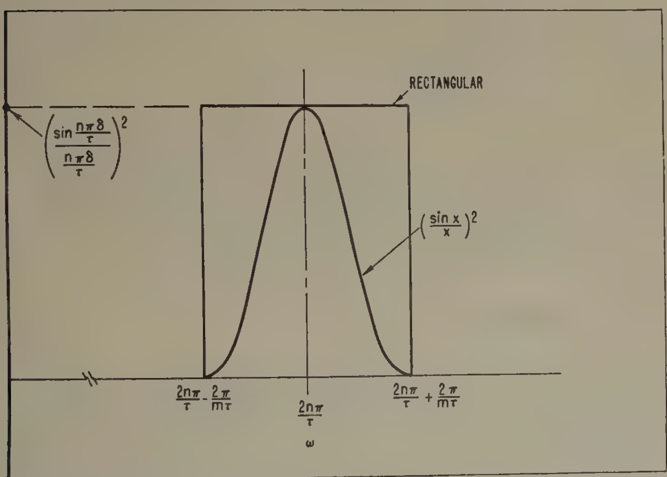

Fig. 8—Typical section of rectangular and North-type $(\sin x/x)^2$ comb-filter frequency response.

It is seen that the filter bandwidth $4\pi/m\tau$, about each prf line, is determined by the characteristics of the communications system employed; that is, it depends upon m, the number of pulses in a sequence, and τ, the pulse period. In a search-radar system the minimum value of m would be the number of pulses that must be reflected by the target so as to be detected.

Fig. 9 is a plot of that part of (17) showing the improvement in signal-to-noise energy ratio gained by the use of comb filtering as a function of the number of pulses m. For example, if as many as 100 pulses can be observed, and stability permits the type of comb filtering being considered, there can result as much as 17-db improvement in sensitivity. This could mean an increase in radar range by a factor of over 2.5.

B. Type II: Rectangular $(\sin x/x)^2$ Comb Filter

Before discussing this filter, it will be necessary to obtain an expression for the unfiltered input signal energy. Since it has previously been established that 90 per cent of the energy of the sin x/x pattern lies in the main hump, the total input energy in the interval $-(2\pi/\delta) \leqq \omega \leqq (2\pi/\delta)$ can be expressed as

$$E = \frac{2\pi\Lambda}{0.90} \sum_{n=-N}^{N} \left[\frac{\sin\dfrac{n\pi\delta}{\tau}}{\dfrac{n\pi\delta}{\tau}}\right]^2 \cdot \int_{2n\pi/\tau-2\pi/m\tau}^{2n\pi/\tau+2\pi/m\tau} \left[\frac{\sin\dfrac{m\tau}{2}\left(\omega-\dfrac{2n\pi}{\tau}\right)}{\dfrac{m\tau}{2}\left(\omega-\dfrac{2n\pi}{\tau}\right)}\right]^2 d\omega \qquad (18)$$

where N is the largest integer in τ/δ and $\Lambda = A^2\delta^2 m^2$. For large N this can accurately be evaluated by approximation methods to be:

$$E = 2\pi \frac{2.84\Lambda\left(2\dfrac{\tau}{\delta}+1\right)}{m\tau}. \qquad (19)$$

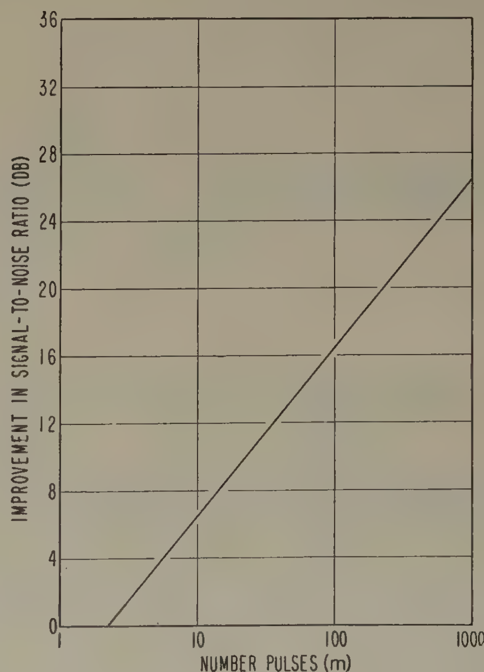

Fig. 9—Improvement in signal-to-noise ratio due to a uniform comb filter as a function of the number of pulses in a sequence.

Now the Type II filter still is an idealized filter in that each separate portion of the frequency characteristic has infinitely steep sides and flat pass band. The basic concept as established in Type I holds here except that instead of all filters having the same gain, they follow the envelope in gain variation. Hence, the gain of the filter about the nth prf line is proportional to the maximum amplitude of the signal spectrum at $\omega = 2n\pi/\tau$ (see Fig. 5).

The signal-energy density about the nth prf line is equal to

$$E_n = \Lambda \left[\frac{\sin \frac{n\pi\delta}{\tau}}{\frac{n\pi\delta}{\tau}}\right]^2 \left[\frac{\sin \frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}{\frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}\right]^2. \quad (20)$$

The output of that element of the comb filter about the nth prf line in the interval

$$\frac{2n\pi}{\tau} - \frac{2\pi}{m\tau} \leqq \omega \leqq \frac{2n\pi}{\tau} + \frac{2\pi}{m\tau}$$

is the area under the energy-density curve given by

$$E_{nCF} = \left[\frac{\sin \frac{n\pi\delta}{\tau}}{\frac{n\pi\delta}{\tau}}\right]^2 E_n. \quad (21)$$

Hence for the Type II comb filter the complete energy output in the interval $-(2\pi/\delta) \leqq \omega \leqq (2\pi/\delta)$ is

$$E_{CF} = 2\pi a\Lambda \sum_{n=-N}^{N} \left[\frac{\sin \frac{n\pi\delta}{\tau}}{\frac{n\pi\delta}{\tau}}\right]^4 \cdot \int_{2n\pi/\tau - 2\pi/m\tau}^{2n\pi/\tau + 2\pi/m\tau} \left[\frac{\sin \frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}{\frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}\right]^2 d\omega. \quad (22)$$

This becomes

$$E_{CF} = 2\pi \frac{1.88\Lambda\left(2\frac{\tau}{\delta} + 1\right)a}{m\tau}. \quad (23)$$

The noise output of this filter is similarly

$$N_{CF} = a\frac{4\pi}{m\tau} N_u \sum_{n=-N}^{N} \left[\frac{\sin \frac{n\pi\delta}{\tau}}{\frac{n\pi\delta}{\tau}}\right]^2 = \frac{0.90 N_o a}{m}. \quad (24)$$

By expressing (23) in terms of (19) the output signal-to-noise ratio for a Type II comb filter is found to be

$$r_{oII} = 10 \log \frac{E_{CF}}{N_{CF}} = r_i + 10 \log 0.45m + 2.1 \text{ db}. \quad (25)$$

This shows that the Type II comb filter will yield only a 2.1-db improvement over the Type I filter.

C. Type III: North-Type $(\sin x/x)^2$ Comb Filter

This final filter is based on the North matched-filter technique and hence should represent the best that can be done with comb filters of the kind discussed here. The derivation of an expression for the signal-to-noise energy ratio follows the same form as for Type II. The signal output is

$$E_{CF} = 2\pi a\Lambda \sum_{n=-N}^{N} \left[\frac{\sin \frac{n\pi\delta}{\tau}}{\frac{n\pi\delta}{\tau}}\right]^4 \cdot \int_{2n\pi/\tau - 2\pi/m\tau}^{2n\pi/\tau + 2\pi/m\tau} \left[\frac{\sin \frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}{\frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}\right]^4 d\omega. \quad (26)$$

This becomes

$$E_{CF} = 2\pi \frac{1.38\Lambda\left(2\frac{\tau}{\delta} + 1\right)a}{m\tau}. \quad (27)$$

Similarly the noise out is

$$N_{CF} = a \sum_{n=-N}^{N} \left[\frac{\sin \frac{n\pi\delta}{\tau}}{\frac{n\pi\delta}{\tau}}\right]^2 \cdot \int_{2n\pi/\tau - 2\pi/m\tau}^{2n\pi/\tau + 2\pi/m\tau} \left[\frac{\sin \frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}{\frac{m\tau}{2}\left(\omega - \frac{2n\pi}{\tau}\right)}\right]^2 d\omega$$

$$= \frac{0.406 N_o a}{m}. \quad (28)$$

Thus the output signal-to-noise energy ratio for the Type III comb filter is

$$\begin{aligned} r_{oIII} &= r_i + 10 \log 1.20m \\ &= r_i + 10 \log 0.45m + 4.3 \text{ db}. \end{aligned} \quad (29)$$

Here again we have the same characteristic as in the other types of filters, except for the 4.3-db fixed improvement over the Type I filter.

Discussion

Up to this point retention of pulse shape has not been mentioned. The signal-to-noise ratios obtained were based on total signal energy and mean noise energy. If pulse shape is important it should be noted that none of the comb filters will reproduce the input pulse shape exactly. The Type I filter will have the least effect on pulse shape, rounding it off somewhat, whereas the Type III filter will alter the pulse shape rather significantly.

In fact, the Type III filter, being very nearly a cross-correlation device, will produce an output which is almost triangular in shape. In some cases this property might be used to advantage.

One rather severe condition has been imposed upon three types of comb filters discussed here. This is that the bandwidth of the individual filter elements is a function of the number of pulses received in a sequence. This presupposes that the receiver must have this prior knowledge and that the number of pulses in the train is fixed. In practice this is seldom true and hence a compromise is required. The formulas for improvement in signal-to-noise ratio are obtained on the premise that as the number of received pulses varies, the filter bandwidths vary correspondingly.

Conclusion

The analysis of a finite sequence of pulses shows that the continuous frequency spectrum consists of a set of sin x/x curves centered about the prf lines; i.e., where the prf lines would be for infinite pulse train. Hence the technique of comb filtering is feasible. For the simplest comb filter, the uniform type tailored to match the number of input pulses in a train, the improvement in signal-to-noise energy ratio is 10 log 0.45 m for m pulses. The comb filter based on the North matched-filter technique yields a constant added improvement of 4.3 db. It appears that such a small gain in sensitivity is hardly justification for the increased complexity in the design and construction of an exact North filter. In any case, although filter elements with infinite slopes are used for mathematical simplicity, in practice the frequency response of a simple uniform filter would more nearly approach the rounded peaks and sloping sides of the North-type filter.

Acknowledgment

Much of the fundamental work presented in the first part of this paper was done by A. G. Davis, now instructor of mathematics at the University of Massachusetts, Amherst. The authors wish to thank C. H. Chrisman, Jr. and his staff in the Operational Research Branch, Radio Division III, NRL, for their valuable assistance in computing the frequency spectra.

Effect of Attenuation on the Choice of Wavelength for Weather Detection by Radar*

WALTER HITSCHFELD† AND J. S. MARSHALL†

Summary—**The performance of radars in storm detection is studied in a manner designed to help in the choice of the proper wavelength for the radar. Decreasing the wavelength increases the range in clear air, but the attenuation associated with the shortened wavelength cuts down appreciably the ability of the radar to look through intervening rain. It is concluded that for the certain detection of operationally significant storms, wavelengths shorter than about 5 cm should not be used.**

Introduction

The performance of a weather radar may be considered under headings of resolution, sensitivity, and accuracy of rain measurement. Decreasing the wavelength would increase resolution and sensitivity if it were not for an accompanying increase in attenuation by rain. The attenuation works against the enhanced resolution by introducing distortion, nearby precipitation patterns casting shadows on more distant patterns. It works powerfully against the enhanced sensitivity; at shorter wavelengths sensitivity depends more on the amount of intervening rain than it does on the distance away from the target rain. Attenuation at the shorter wavelengths also renders rain measurement by radar next to impossible, as the necessary correcting procedure is difficult to apply.[1] The present paper will not deal with resolution and measurement, but will deal carefully with the matter of sensitivity.

Theory of Radar Performance

Provided the beam is filled with scatterers and provided that attenuation due to intervening rain and

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[1] W. Hitschfeld and J. Bordan, "Errors inherent in the radar measurement of rainfall at attenuating wavelengths," *Jour. Met.*, vol. 10, pp. 58–67; February, 1954.

atmospheric gases may be neglected, the performance of a weather radar may be described by the equation

$$P_r = \frac{P_0 h A \sum \sigma}{8\pi r^2}\left(\frac{A_e}{A} F\right), \tag{1}$$

where P_r is the power received at the radar, P_0 is the power transmitted, h is the pulse length, A is the geometrical area of the antenna, $\Sigma\sigma$ is the sum of the back-scatter cross sections of all the precipitation particles in unit volume, and r is the range. A_e is the effective area of the antenna for extended targets and is less than A. The exact value of A_e is not easily determined. For this reason A only is retained in the equation and a conservative estimate ($\frac{1}{2}$) is made for A_e/A. An equation identical to (1), but without F, is derived by Kerr.[2] The factor F, due to imperfectly known causes, is inserted on empirical grounds. Austin and Williams[3] have shown F to be about 1/5 for their SCR-615-B radar. In the absence of general information, it is considered best to adopt this value, so that a value 0.10 results for the factors in the bracket. It is felt that with these factors, (1) represents the performance of a well-maintained radar as closely as present knowledge permits.

The back-scatter cross section of a raindrop is given, according to Rayleigh's approximation, as

$$\sigma = \frac{\pi^5 D^6 |\kappa|^2}{\lambda^4}, \tag{2}$$

where D is the drop diameter, and

$$\kappa = \frac{m^2 - 1}{m^2 + 2},$$

m being the complex refractive index of water, and λ the wavelength of the radar.[4]

The sum of the sixth powers of the drop diameters in unit volume is given by

$$\sum D^6 = 2.0 R^{1.6}\, 10^{-10} \text{ cm}^6 \text{ cm}^{-3}, \tag{3}$$

where R is the rate of rainfall in mm hr^{-1}. (Equation (3) is a slight improvement on a similar equation containing the numerical factor 2.2×10^{-10} reported by Marshall and Palmer.[5])

Combining (1), (2), and (3), and solving for R one gets:

$$R = \lambda^{2.50}\, r^{1.25} \cdot P_r^{0.625} \left[25\pi^4\, 10^{-13} A\, |\kappa|^2 P_0\right]^{-0.625}, \tag{4}$$

with all the quantities in cgs units, and R in mm hr^{-1}. (Here $(A_e/A)F$ has been put equal to 0.10 as discussed above.)

The minimum detectable rainfall R_m may be calculated from this equation if one uses the minimum detectable signal power for P_r. At the 1952 Radar Weather Conference such minimum values were reviewed (see Appendix I), and the consensus of opinion seemed to be that minimum detectable $P_r (= P_{rm})$ is between 2 and 4×10^{-13} watts for 3.2-, 5.7-, and 10-cm radar. A value of P_{rm} of 4×10^{-13} watts will be used throughout this paper.

Putting $|\kappa|^2 = 0.93$ (a value very nearly independent of wavelength) re-write (4) for the minimum detectable rainfall in more convenient units:

$$\underset{(\text{in mm hr}^{-1})}{R_m} = 0.0981 \left(\frac{\lambda}{5.7 \text{ cm}}\right)^{2.50} \left(\frac{r}{\text{in mi}}\right)^{1.25} \left(\frac{P_{rm}}{4 \times 10^{-13} \text{ w}}\right)^{0.625} \cdot \left(\frac{P_0}{40 \text{ kw}} \frac{h}{300 \text{ m}}\right)^{-0.625} \left(\frac{d}{18''}\right)^{-1.25} \tag{5}$$

where d is the diameter of the (circular) antenna.

When the beam is not filled by precipitation at range r, the equation needs to be modified. The diameter of a circular beam is (between half-power points) $r(1.14\lambda/d)$; hence, if we view a storm of linear dimension L, it follows that the beam will be just filled at range r_0, such that

$$r_0 \frac{1.14\lambda}{d} = L. \tag{6}$$

Beyond r_0 the beam will no longer be filled, and here P_r in (1) should be proportional to r^{-4} rather than to r^{-2}. Carrying through this modification, taking care that the two forms of (1) must agree at $r = r_0$, results in

$$\underset{\text{in mm hr}^{-1}}{R_m'} = 0.00218 \left(\frac{\lambda}{5.7 \text{ cm}}\right)^{3.75} \left(\frac{r}{\text{in m}}\right)^{2.50} \left(\frac{P_{rm}}{4.10^{-13} \text{ w}}\right)^{.625} \cdot \left(\frac{P_0}{40 \text{ kw}} \frac{h}{300 \text{ m}}\right)^{-.625} \left(\frac{d}{18''}\right)^{-2.50} \left(\frac{L}{3 \text{ mi}}\right)^{-1.25}. \tag{7}$$

Equations (5) and (7) give the minimum detectable rainfall without allowing for attenuation by intervening rain and by air and water vapor. Such attenuation causes a drop in the received power, and may be allowed for most easily by a corresponding drop in P_0, or an equivalent increase in $R_m^{1.6}$, in the above equations.

Attenuation by rain may be expressed accurately by:

$$\underset{(\text{in db})}{\text{radar attenuation by rain}} = K' \int_0^r R^\alpha dr. \tag{8}$$

Values of the constants K' and α are listed in Table I.

TABLE I

ATTENUATION BY RAIN AND ATMOSPHERIC GASES

λ[cm]	3.2	5.7	10
K'db(mm hr^{-1})$^{-\alpha}$mi$^{-\alpha}$]	0.0288	0.0094	0.00098
α	1.3	1.1	1.0
K[db(mm hr^{-1} mi)$^{-1}$]	0.08	0.013	0.00098
k[db mi^{-1}]	0.054	0.036	0.013

These values are taken largely from Ryde.[6]

[2] D. E. Kerr, Ed., "Propagation of Short Radio Waves," vol. 13, Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y.; 1951.

[3] P. M. Austin, and E. L. Williams, Jr., "Comparison of Radar Signal Intensity with Precipitation Rate," Technical Report No. 14, submitted by Massachusetts Institute of Technology to Signal Corps Engineering Laboratories; June 1, 1951.

[4] Kerr, *ibid.*, sec. 7.3.

[5] J. S. Marshall and W. M. Palmer, "The distribution of raindrops with size," *Jour Met.*, vol. 5, pp. 165–166; August, 1948.

The values of α are sufficiently close to unity to allow the simplification:

radar attenuation by rain
(in db)

$$= K \times \text{quantity of intervening rain,} \qquad (9)$$
$$(\text{in mm hr}^{-1} \times \text{miles})$$

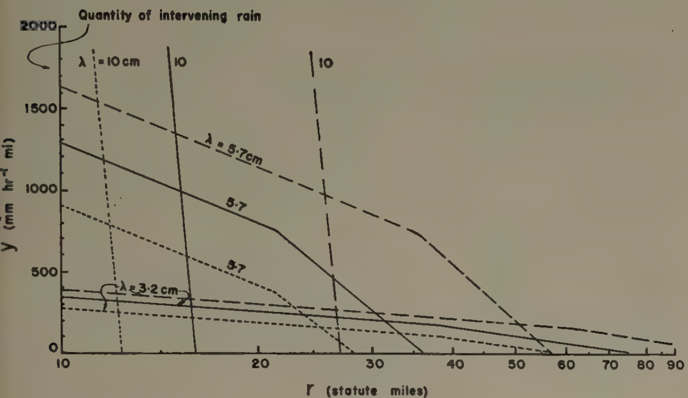

Fig. 1—Showing the amount of intervening rain through which a target rainfall (R_0) of dimensions 3 miles $\times$ 3 miles and at range r may be detected.
Solid lines: $P_0=120$ kw, $d=18$ inches, $R_0=10$ mm hr^{-1}
Broken lines: $P_0=120$ kw, $d=30$ inches, $R_0=10$ mm hr^{-1}
Dotted lines: $\begin{cases} P_0=40 \text{ kw}, d=18 \text{ inches}, R_0=10 \text{ mm hr}^{-1} \text{ or} \\ P_0=120 \text{ kw}, d=18 \text{ inches}, R_0=5 \text{ mm hr}^{-1}. \end{cases}$
Kinks of the solid and broken 10-cm loci are off the diagram at ranges 11.8 and 19 miles respectively.

where K is only slightly dependent on R. Values of K corresponding to R about 30 mm hr^{-1}, are also listed in Table I.

The attenuation by atmospheric gases (notably oxygen and water vapor) varies with temperature, pressure, and humidity. Since the values are small, only representative figures corresponding to a temperature just above 0°C, standard pressure, and water vapor pressure close to saturation, need be used. Values of k, the radar attenuation by gases, are shown in Table I.

Graphical Display of Performance

A practically useful way of portraying the effect of the many parameters involved is to plot the intervening rain y (in mm hr^{-1} miles), through which a storm of given extent and intensity may be detected, as a function of range r. It is easy to show that in the absence of gas attenuation, y is given by

$$y = -\frac{20}{K}\log r + \frac{16}{K}\log\frac{R_0}{C} \qquad (10)$$

or by

$$y = -\frac{40}{K}\log r + \frac{16}{K}\log\frac{R_0}{C'}, \qquad (11)$$

depending on whether or not the beam is filled with target. Here, R_0 is the rainfall that can actually be detected at range r, and $C=R_m/r^{1.25}$ and $C'=R_m'/r^{2.50}$ are two constants depending only on the radar parameters which can be worked out from (5) and (7). Plots of y and y' against log r are straight lines. Allowance for gas attenuation is easily made, simply by reducing, for any value of r, the value of y or y' by an amount

$$\Delta y = \Delta y' = \frac{kr}{K}. \qquad (12)$$

This correction is quite small, so that in spite of it, the corrected loci of y and y' are still approximately straight. The emphasis in this paper is on the effect of wavelength. Loci of y or y' were therefore drawn in groups of three: one locus for 3.2, one for 5.7, and one for 10 cm, with all other parameters unchanged.

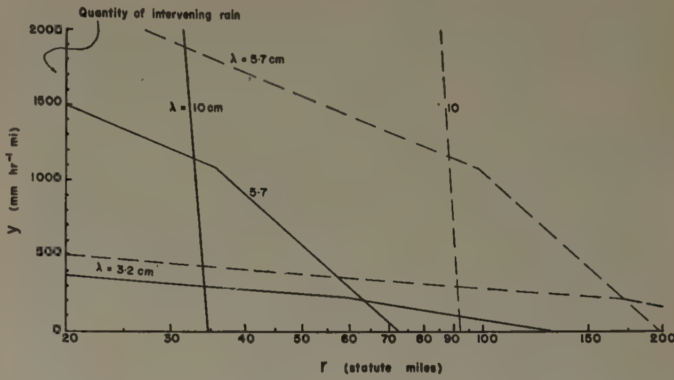

Fig. 2—Showing the amount of intervening rain through which a target rainfall of intensity 10 mm hr^{-1}, of dimensions 3 miles $\times$3 miles, and at range r may be detected.
Solid lines: $P_0=350$ kw, $d=30$ inches
Broken lines: $P_0=350$ kw, $d=82.5$ inches.
(350 kw and 82.5 inches are characteristics of the AN/CPS-9 radar.) Kinks in the 10-cm loci occur off the diagram at 19 and 52 miles.

Discussion

The results are in Figs. 1 and 2, above. The graphs of Fig. 1 are for radars having values of power and size of antenna suitable for airborne equipment. The solid lines, for instance, refer to transmitter power 120 kw, antenna aperture 18 inches diameter, pulse length 300 meters, intensity of target rainfall 10 mm hr^{-1}. For amounts of intervening rain up to 200 mm hr^{-1} miles, the 3.2-cm equipment gives greater ranges than the 5.7 cm. With this much intervening rain, both wavelengths have a limiting range of 31 miles. (The 10-cm equipment can cope with almost any amount of intervening rainfall, but its short range and very low angular resolution (not further brought out here) make it an unlikely choice for use with these values of power and size of antenna.) As one proceeds to greater amounts of intervening rain, the range at 3.2 cm continues to drop off very rapidly, so that with twice this much intervening rain the range is about 7 miles. At 5.7 cm, on the other hand, the same 400 mm hr^{-1} miles of rain drops the range back by less than 15 per cent to 27 miles. It was necessary to specify the extent of the target shower, since this determines whether or not the target precipitation completely fills the cross section of the radar beam. A shower of linear dimensions 3 miles high by 3

[6] J. W. Ryde, Contribution in "Meteorological factors in radio wave propagation," Phys. Soc., (London, Eng.); 1946.

miles wide was chosen rather arbitrarily for all curves. At 5.7 cm this target fills the beam of the 18-inch antenna at ranges up to 21 miles, the narrower 3.2-cm beam at ranges up to 38 miles. Kinks in the loci under discussion may be noted at those ranges. The kink on the 10-cm locus is off the diagram, at about 11.8 miles.

Fig. 2 shows the results of similar calculations using values of power and size of antenna more appropriate to ground-based equipment. The advantage of using a large antenna is seen to be considerable, and to increase as the wavelength is increased.

The two charts show five sets of three curves, covering six sets of parameters at the three wavelengths. It is, however, easy to construct further approximate loci by moving the lines, without changing their direction, in the following way. If instead of P_0, the transmitter power is P_0', the curves should be moved *vertically* by changing the height of any point at fixed range by $(10/K) \log P_0'/P_0$. Similarly, if the pulse length is changed from h to h', or the target rainfall from R_0 to R_0', the vertical shifts should be $(10/K) \log h'/h$, or $(16/K) \log R_0'/R_0$, respectively. If instead of d, we have an antenna of diameter d', the curves should be moved *horizontally*, changing the range of any point of given y from an initial value of r to a final one r' according to $r'/r = d'/d$. The effect of changing the linear dimension of target rain from L to L' is to move the kink along the left part of the locus (produced if necessary) from a range r_0 to a new range r_0' according to $r_0'/r_0 = L'/L$. The right part of the locus is then drawn parallel to the right part of the original locus.

Conclusion

Attenuation by intervening rainfall effects a very serious limitation to the sensitivity of 3-cm weather radar. A frontal line of showers is likely to involve the passage of the radar beam through a few hundred mm hr^{-1} miles of rain; probably, though not certainly, less than 500. Sensitivity at 3 cm varies very rapidly with the amount of intervening rain, and this rapid variation in itself would lead to uncertainty in interpreting signals. Five hundred mm hr^{-1} miles is enough to render 3-cm equipment practically inoperative. At wavelength 5.7-cm, attenuation is appreciable but the sensitivity changes more gradually with intervening rain, and the range would never be reduced by more than 30 per cent. At this wavelength, however, it is difficult for small radars to achieve even in clear air the ranges which may reasonably be desired. It should be noted however that the range increases with the intensity of target rainfall. The range is doubled in going from the 10 mm hr^{-1} illustrated here to 60 mm hr^{-1}. Radar observations on thundershowers usually indicate intensities of this order at the core, or anyway a core condition providing a signal equivalent to 60 mm hr^{-1} in radar reflectivity.

The three wavelengths have been compared only in sensitivity. For a constant size of antenna, they will also differ in resolution: the shorter wavelength will have a correspondingly narrower beam, correspondingly higher resolution. But attenuation will introduce distortion into the pictures, altering outlines and casting shadows of nearer showers on more distant ones. It may be that the initially better resolution of the shorter wavelength will be more than cancelled out by attenuation, in the same way as the initially higher sensitivity. The situation with regard to resolution and distortion has not been studied; the best reference to date would appear to be Atlas and Banks.[7]

For the powers and antenna sizes available to ground-based weather radar, Fig. 2 indicates that each of the three wavelengths studied has its virtues. At 3.2 cm the resolution will be best, apart from distortion by attenuation. That is to say, resolution on the leading edge of the storm area will be best and with a large antenna, the radiation can penetrate a fair amount of precipitation. The range at 5.7 cm is over 100 miles through any likely amount of intervening rain: there is thus no longer doubt of achieving satisfactory range at wavelength 5.7 cm. Again with the large antenna, the 10-cm radiation reaches to the useful range of 90 miles, without the distortion and error in rain measurement resulting from attenuation.

It may be worth noting that the considerations of this paper do not apply when the precipitation is in the form of dry snow. In that case attenuation is negligible at all wavelengths, so that full benefit is obtained of the increase of scattering and improved resolution when the wavelength is reduced. Such an increase in sensitivity is important, for ice scatters less well than water by a factor five, and the intensity of precipitation to be observed as snow is much more likely to be 1 mm (of water) per hour than the 10 mm hr^{-1} considered here.[8]

Appendix I

Power Characteristics of Radars of Different Wavelengths

The following table is taken from a brief note (pages B-37/38) in the Proceedings of the Third Radar Weather Conference, held at McGill University, September 17 to 19, 1952. The data represent the general consensus of the Conference, which comprised the majority of radar weather specialists.

Table II

λ(cm)	0.9	1.25	3.2	5.6	10.0
Maximum aecvailable transmitter power (P_0, *kw*)	20	40	350	300	1,000
Minimum receivable signal power (P_{rm} watts)	10^{-12}	5×10^{-13}	4×10^{-13}	3×10^{-13}	2×10^{-13}

[7] D. Atlas and H. C. Banks, "The interpretation of microwave reflections from rainfall," *Jour. Met.*, vol. 8, pp. 271–282; October, 1951.

[8] J. S. Marshall and K. L. S. Gunn, "Measurement of snow parameters by radar," *Jour. Met.*, vol. 9, pp. 322–327; October, 1952.

Generalized Equations for RC Phase-Shift Oscillators*

SOL SHERR†, SENIOR MEMBER, IRE

Summary—General solutions are obtained for three- and four-section phase-shift networks. These solutions are reduced to design equations for a number of specific circuit configurations, and the complete solution is given for a four-section phase-shift oscillator, including effects of plate-load resistance, and input-miller capacity.

RC PHASE-SHIFT OSCILLATORS exhibit several useful characteristics, foremost among which are cost, weight, and space reduction. The frequency-determining components of these oscillators are relatively tiny and inexpensive in comparison to the coils and capacitors of conventional resonant circuits, the difference being particularly striking in the low audio-frequency region. This type of oscillator is being used with increasing frequency by designers who wish the advantage of reduced weight and cost while maintaining good waveform and stability in the audio range.

The subject of this class of oscillators has been treated rather extensively in the literature.[1–6] However, the treatment has usually been in terms of a single type or restricted application, with the resultant lack of generality. This article proposes to develop completely generalized formulas for phase-shift oscillators, then reduce these formulas to various specific cases, some well known and some new. The analysis will be restricted, for practical reasons, to three- and four-section networks.

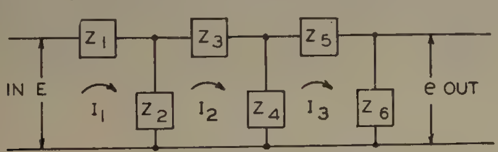

Fig. 1—Generalized three-mesh network.

The minimum number of rc sections with which it is possible to achieve oscillation is three. This network is shown in Fig. 1.

Solving for I_3, we have

$$I_3 = \frac{EZ_2Z_4}{(Z_4+Z_5+Z_6)(Z_1+Z_2)(Z_2+Z_3+Z_4) - Z_2^2(Z_4+Z_5+Z_6) - Z_4^2(Z_1+Z_2)}$$

and $e = I_3Z_6$.

The transfer characteristic is defined by

$$\frac{e}{E} = \frac{Z_2Z_4Z_6}{(Z_1+Z_2)(Z_2+Z_3+Z_4)(Z_4+Z_5+Z_6) - Z_2^2(Z_4+Z_5+Z_6) - Z_4^2(Z_1+Z_2)}. \tag{1}$$

This equation defines the frequency of oscillation and the gain required to produce oscillation. The frequency of oscillation, "F" is determined by setting the imaginary, or odd-order terms equal to zero, which satisfies the condition that the result be a negative real quantity. The required gain, "A" is found by substituting the values of Z, derived from the frequency solution, into the real part of the transfer characteristic and solving for the loss of the network. This loss is what must be compensated for by external gain in order to satisfy the requirement that the entire circuit be lossless.

If the substitutions $Z_1=Z_3=Z_5$, and $Z_2=Z_4=Z_6$ are made, the general equation reduces to

$$\frac{e}{E} = \frac{Z_2^3}{Z_2^3 + 6Z_2^2Z_1 + 5Z_1^2Z_2 + Z_1^3} \tag{2}$$

which is the one derived by Gamertsfelder.[7] Further substitutions of $Z_1 = -jx_c$ and $Z_2 = R$, will give the results

$$F = \frac{1}{2\pi RC\sqrt{6}} \tag{2A}$$

$$A = -29, \tag{2B}$$

which are the ones derived by Ginzton and Hollingsworth.[2] Similarly letting $Z_1 = R$ and $Z_2 = -jx_c$ will give

$$F = \frac{\sqrt{6}}{2\pi RC}, \tag{2C}$$

which was derived by Ginzton and Hollingsworth[2] and

$$A = -29, \tag{2D}$$

which is the correction made by Blanchard[3] of the Ginzton and Hollingsworth result of $A=-5$.

A further substitution of $Z_3 = KZ_1$, $Z_5 = K^2Z_1$, $Z = KZ_2$, $Z_6 = K^2Z_2$ results in the general formula

$$\frac{e}{E} = \frac{1}{\dfrac{Z_1^3}{Z_2^3} + \dfrac{(3K+2)}{K}\cdot\dfrac{Z_1^2}{Z_2^2} + \dfrac{(3K^2+2K+1)}{K^2}\cdot\dfrac{Z_1}{Z_2} + 1}. \tag{3}$$

This network is shown in Fig. 2. The specific results derived by Johnson,[4] namely

* Decimal classification: R355.914.31. Original manuscript received by the IRE, September 17, 1953; revised manuscript received, February 10, 1954.

† General Precision Laboratory Inc., Pleasantville, N. Y.

[1] Britton Chance et al., "Waveforms," McGraw-Hill Book Co., New York, N. Y.; 1949.

[2] E. L. Ginzton and L. M. Hollingsworth, "Phase shift oscillators," Proc. I.R.E., vol. 29, pp. 43–49; February, 1941.

[3] A. Blanchard, "Correction to phase shift oscillators by Ginzton and Hollingsworth," Proc. I.R.E., vol. 32, p. 641; October, 1944.

[4] R. W. Johnson, "Extending the frequency range of the phase shift oscillator," Proc. I.R.E., vol. 33, pp. 597–603; September, 1945.

[5] P. G. Sulzer, "The tapered phase shift oscillator," Proc. I.R.E., vol. 36, pp. 1302–1305. October, 1948.

[6] W. C. Vaughan, "Phase shift oscillator," *Wireless Eng.*, vol. 26, pp. 391–399; December, 1949.

[7] B. Chance, *op. cit.*, p. 111.

$$F = \frac{1}{2\pi RC\sqrt{3 + \dfrac{2}{K} + \dfrac{1}{K^2} + \left(2\dfrac{R_1}{R}\right)\left(\dfrac{1}{K} + 1\right)}} \tag{3A}$$

and

$$A = -8 - \frac{R_1}{R}\left[\frac{11}{K} + \frac{4}{K^2} + 8\right] - \left(\frac{R_1}{R}\right)^2\left(\frac{2}{K} + 2\right), \tag{3B}$$

can also be found from (1) by making the appropriate substitutions

$$Z_1 = (R_1 - jx_c), \qquad Z_3 = Z_5 = -jx_c,$$
$$Z_2 = Z_4 = Z_6 = R.$$

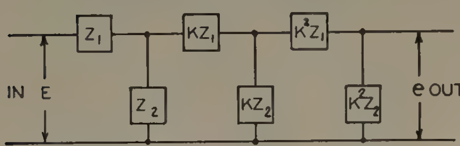

Fig. 2—Tapered three-mesh network.

Any other desired combination of impedances may be used and the specific formulas derived from the generalized one given above. For example let $Z_2 = Z_4 = Z_6$, and $Z_3 = Z_5$. The network is shown in Fig. 3. The equation becomes

$$\frac{e}{E} = \frac{1}{3\dfrac{Z_1}{Z_2} + 4\dfrac{Z_1Z_3}{Z_2^2} + \dfrac{Z_1Z_3^2}{Z_2^3} + 1 + 3\dfrac{Z_3}{Z_2} + \dfrac{Z_3^2}{Z_2^2}}. \tag{4}$$

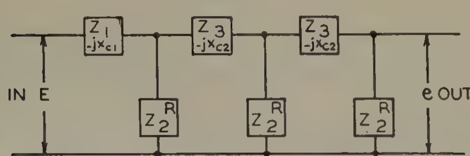

Fig. 3—Shunt R three-mesh network.

Equating the odd-order terms to zero, substituting

$$Z_2 = Z_4 = Z_6 = R,$$
$$Z_1 = -jXc_1,$$

and

$$Z_3 = Z_5 = -jXc_2,$$

and simplifying, gives

$$F = \frac{1}{2\pi R\sqrt{3(C_2^2 + C_1C_2)}}, \tag{5}$$

which is equivalent to the result given by Ginzton and Hollingsworth.[2] Formulas for any combinations of impedances may be derived by proper substitution into the general formula.

The general four-mesh network is shown in Fig. 4. Solving for i_4, we have

$$i_4 = \frac{EZ_2Z_4Z_6}{a(Z_2+Z_3+Z_4)(Z_1+Z_2) - Z_2^2a - (Z_1+Z_2)(Z_6+Z_7+Z_8)}$$

where

$$a = (Z_6 + Z_7 + Z_8)(Z_4 + Z_5 + Z_6) - Z_6^2,$$

and the transfer characteristic is defined by

$$\frac{e}{E} = \frac{1}{\dfrac{(bcde')}{Z_2Z_4Z_6Z_8} - \dfrac{Z_6(de')}{Z_2Z_4Z_8} - \dfrac{Z_4(be')}{Z_2Z_6Z_8} - \dfrac{Z_2(bc)}{Z_4Z_6Z_8} + \dfrac{Z_2Z_6}{Z_4Z_8}} \tag{6}$$

where

$$b = (Z_6 + Z_7 + Z_8)$$
$$c = (Z_4 + Z_5 + Z_6)$$
$$d = (Z_2 + Z_3 + Z_4)$$

and

$$e' = (Z_1 + Z_2).$$

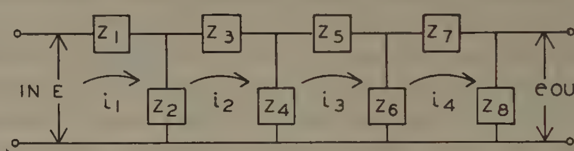

Fig. 4—Generalized four-mesh network.

This equation defines the frequency of oscillation "F" and the gain "A" necessary for oscillation in the same manner as (1) does for the three-section network.

If

$$Z_1 = Z_3 = Z_5 = Z_7,$$

and

$$Z_2 = Z_4 = Z_6 = Z_8,$$

then

$$\frac{e}{E} = \frac{1}{\dfrac{(f^2g)}{Z_2^4} - \dfrac{2Z_2(fg)}{Z_2^3} - \dfrac{Z_2f^2}{Z_2^3} + 1} \tag{7}$$

where

$$f = (Z_1 + 2Z_2)$$
$$g = (Z_1 + Z_2).$$

Let $K_0 = Z_1/Z_2$ and simplify, then

$$\frac{e}{E} = \frac{1}{K_0^4 + 7K_0^3 + 15K_0^2 + 10K_0 + 1} \tag{8}$$

and equating odd-order terms to zero gives

$$K_0 = \sqrt{\frac{10}{7}}.$$

Let

$$Z_1 = -jX_c,$$

and

$$Z_2 = R,$$

then

$$F = \frac{1}{2\pi CR\sqrt{\dfrac{10}{7}}} \tag{9}$$

which is the formula given by Ginzton and Hollingsworth.[2] Substituting for K_0 in the real terms, we have the result,

$$A = \frac{100}{49} - \frac{150}{7} + 1 = -18.4 \tag{9A}$$

as the minimum gain required for oscillation. This corresponds to the correction of the Ginzton and Hollingsworth[2] result given by Blanchard.[3] In addition, if the substitutions $Z_1 = Z_3 = -jX_c$, $Z_5 = Z_7 = -jX_{c1}$, and $Z_2 = Z_4 = Z_6 = Z_8 = R$ are made, then the final correction by Blanchard,[3] which is

$$A = \frac{9\left(\dfrac{C}{C_1}\right)^3 + 114\left(\dfrac{C}{C_1}\right)^2 + 352\left(\dfrac{C}{C_1}\right) + 342 + 84\left(\dfrac{C_1}{C}\right)}{\left(4 + 3\dfrac{C}{C_1}\right)} \tag{9B}$$

can be derived.

If we wish to find the effect of the vacuum tube in the complete circuit shown in Fig. 5(a) we may use the equivalent circuit of Fig. 5(b) and make the following substitutions:

$$Z_1 = R' + Z_1',$$

$$Z_2 = Z_4 = Z_6 = Z_8,$$

and

$$Z_3 = Z_5 = Z_7 = Z_1'.$$

Then

$$\frac{E}{e} = \frac{(Z_1' + 2Z_2)^3(Z_1' + Z_2 + R')}{Z_2^4} - \frac{2(Z_1' + 2Z_2)(Z_1' + Z_2 + R')}{Z_2^2} - \frac{(Z_1' + 2Z_2)^2}{Z_2^2} + 1.$$

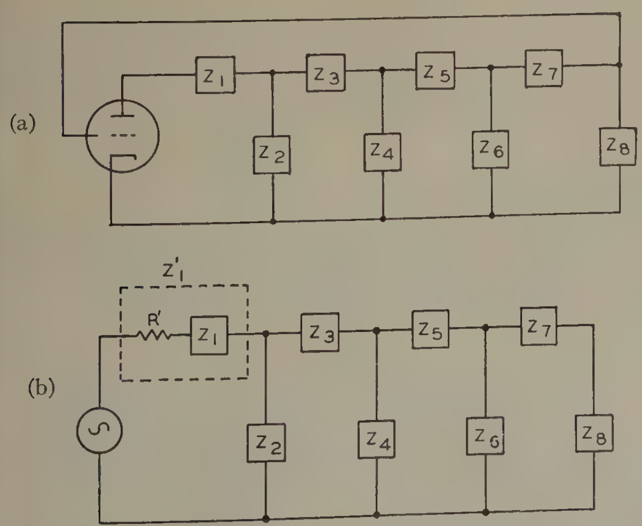

Fig. 5—(a) Vacuum tube with generalized four-mesh network. (b) Vacuum tube equivalent circuit with generalized four-mesh network.

Let $Z_2 = R$ and

$$\frac{Z_1'}{Z_2} = K_1,$$

then

$$\frac{E}{e} = K_1^4 + \left(7 + \frac{R'}{R}\right)K_1^3 + \left(15 + 6\frac{R'}{R}\right)K_1^2 + \left(10 + 10\frac{R'}{R}\right)K_1 + 4\frac{R'}{R} + 1.$$

Equating j terms to zero and substituting $Z_1' = -jX_c$, $K_1 = -jX_c/R$ gives

$$F = \frac{1}{2\pi CR\sqrt{\dfrac{10R\left(1 + \dfrac{R'}{R}\right)}{7R + R'}}}. \tag{10}$$

If we wish to assess the effect of input and Miller capacity we let $Z_1 = Z_3 = Z_5 = Z_7$, and $Z_2 = Z_4 = Z_6$.

The equation becomes

$$\frac{E}{e} = \frac{Z_1^4}{Z_2^3 Z_8} + \frac{6Z_1^3}{Z_2^2 Z_8} + \frac{10Z_1^2}{Z_2 Z_8} + \frac{4Z_1}{Z_8} + \frac{Z_1^3}{Z_2^3} + \frac{5Z_1^2}{Z_2^2} + \frac{6Z_1}{Z_2} + 1.$$

Substituting

$$Z_8 = \frac{-jRXc_1}{R - jXc_1},$$

where C_1 = input and Miller capacity,

$$Z_1 = -jX_c,$$

and

$$Z_2 = R,$$

and equating j terms to zero we get

$$F = \frac{1}{2\pi CR\sqrt{\dfrac{(C_1 + C)10}{C_1 + 7C}}}. \tag{11}$$

If $(R_1C_1/RC) \ll 1$, this may be combined with the previous result to give

$$F = \frac{1}{2\pi CR\sqrt{\dfrac{10}{7}\dfrac{\left(1+\dfrac{C_1}{C}+\dfrac{R'}{R}\right)}{\left(1+\dfrac{C_1}{7C}+\dfrac{R'}{7R}\right)}}} \tag{12}$$

which is the complete solution for this circuit.[8] If $(R'C_1/RC) \approx 0.01$, the error in calculation will be less than 1 per cent.

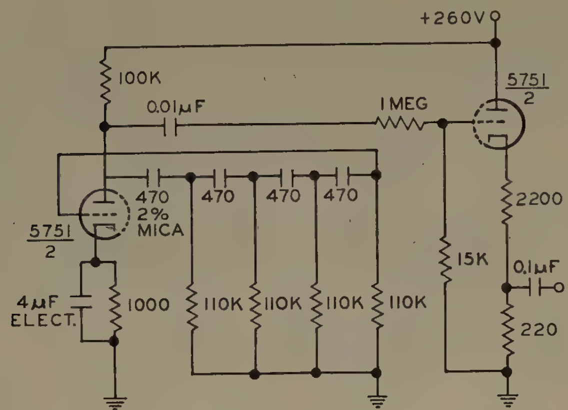

Fig. 6—Test circuit using four-mesh network.

This formula was used to calculate the frequency for the circuit shown in Fig. 6. The values of R' and C_1 were assumed to be approximately 11,000 ohms and 47 MMF respectively. Then

$$F = \frac{1}{6.28\times 470\times 10^{-12}\times 110\times 10^3\sqrt{\dfrac{10}{7}\left[\dfrac{1+.1+.1}{1+\dfrac{1}{70}+\dfrac{1}{70}}\right]}}$$

$\approx$2,300 cycles per second.

The measured frequency differed from the calculated frequency by less than 3 per cent, with the difference attributable to variations in R' and C_1 from the assumed values.

A final example is the tapered four-section network discussed by Sulzer.[5] The substitutions here are $Z_3 = KZ_1$, $Z_4 = KZ_2$, $Z_5 = K^2Z_1$, $Z_6 = K^2Z_2$, $Z_7 = K^3Z_1$ and $Z_8 = K^3Z_2$. Therefore

$$\frac{E}{e} = \frac{(Lmnp)}{K^6Z_2^4} - \frac{K^2Z_2(np)}{K^4Z_2^3} - \frac{KZ_2(Lp)}{K^5Z_2^3} - \frac{Z_2(Lm)}{K^6Z_2^3} + \frac{1}{K^2}.$$

Where

$$L = (K^2Z_2 + K^3Z_1 + K^3Z_2)$$
$$m = (KZ_2 + K^2Z_1 + K^2Z_2)$$
$$n = (Z_2 + KZ_1 + KZ_2)$$
$$p = (Z_1 + Z_2).$$

This reduces to

$$\frac{E}{e} = \frac{Z_1^4}{Z_2^4} + \frac{(4K+3)}{K}\cdot\frac{Z_1^3}{Z_2^3} + \frac{(6K^2+6K+3)}{K^2}\cdot\frac{Z_1^2}{Z_2^2} + \frac{(4K^3+3K^2+2K+1)}{K^3}\cdot\frac{Z_1}{Z_2} + 1.$$

Equating odd terms to zero gives

$$\frac{Z_1}{Z_2} = \sqrt{\frac{4K^3+3K^2+2K+1}{K^2(4K+3)}} \tag{13}$$

and

$$A = \frac{Z_1^4}{Z_2^4} + \frac{(6K^2+6K+3)}{K^2}\cdot\frac{Z_1^2}{Z_2^2} + 1.$$

Substituting for Z_1/Z_2 we have

$$A = \frac{64K^6+192K^5+260K^4+214K^3+119K^2+44K+8}{16K^6+24K^5+9K^4} \tag{14}$$

which corresponds to Sulzer's[5] results, with greater accuracy in the K^2 term. Substituting $K=1$ gives $A=18.4$ as in the simpler case to which it is equivalent, while substituting $K=2$ gives $A=8.65$, which indicates the advantage to be gained from using a tapered network.

Other equations for any combination of resistors and capacitors can be derived from these general equations. Thus, it is possible by direct substitution to solve networks employing these elements.

Acknowledgment

The writer thanks H. Strell for checking the mathematics, R. Bernstein for doing experiments on final circuit, and some calculations. Also E. B. Hales and I. A. Greenwood, Jr. helped offer valuable suggestions.

Appendix I

Equating j terms to zero gives

$$0 = \frac{1}{(\omega CR)^3}\left(7+\frac{C_1}{C}+\frac{R'}{R}\right) - \frac{1}{\omega CR}\left(10+\frac{10C_1}{C}+\frac{10R'}{R}+\frac{5R'C_1}{RC}\right) + \omega C_1R'$$

$$\therefore \quad (\omega CR)^2 = \frac{a \pm \sqrt{a^2 - 4\left(7+\dfrac{C_1}{C}+\dfrac{R'}{R}\right)\left(\dfrac{R'C_1}{RC}\right)}}{2\dfrac{R'C_1}{RC}}$$

where

$$a = 10\left(1+\frac{C_1}{C}+\frac{R'}{R}+\frac{R'C_1}{2RC}\right).$$

Then

$$F = \frac{1}{2\pi CR}\sqrt{\frac{a \pm \left[a^2 - 4\left(7+\dfrac{C_1}{C}+\dfrac{R'}{R}\right)\left(\dfrac{R'C_1}{RC}\right)\right]^{1/2}}{2\dfrac{R'C_1}{RC}}}.$$

The plus solution corresponds to the zero-phase shift case and does not satisfy requirements for oscillation.

[8] See Appendix I for the exact solution.

An Analysis of Passive Reflector Antenna Systems*

R. E. GREENQUIST†, ASSOCIATE, IRE, AND A. J. ORLANDO†

Summary—**The system gain and radiation pattern of a passive reflector antenna system (with a 45-degree plane rectangular reflector) are developed from aperture field theory. Experimentally determined system gains and a measured radiation pattern substantiate the theoretical analysis. Side-lobe levels and mutual coupling are also discussed.**

Introduction

PASSIVE reflector antenna systems have come into rather widespread use during recent years. However, there has been little information on this type antenna system in the literature.

Western Union erected a tower at Monsey Heights, N. Y., to study the performance of such antenna systems for possible use in their microwave radio relay systems. Tests were conducted locally and over the 27.8-mile path to New York City. System gains, mutual coupling, and radiation patterns were measured in the 4,000-mc region.

Most of the theoretical analysis which follows is based upon concepts which are to be found in the literature on microwave antennas and optics. This paper is basically a correlation of these concepts with reference to passive reflector antenna systems. As far as the authors know, this is the first time that experimentally determined gains and radiation patterns have been presented.

Equivalent System

A parabolic antenna may be represented by a uniform-phase circular aperture and a 45-degree plane rectangular reflector may be represented by a 45-degree rectangular aperture with a linear phase error. Fig. 1 shows an equivalent of a passive reflector antenna system. However, in this equivalent system, the 45-degree plane reflector is approximated by a uniform-phase square aperture defined by the projected area of the reflector normal to the incident radiation. This approximation is valid because the aperture is highly directive and the concern is with radiation at angles close to the direction of maximum radiation intensity. In this region there is no significant difference between the radiation characteristics of this aperture and the radiation characteristics of a 45-degree rectangular aperture with a linear phase error.

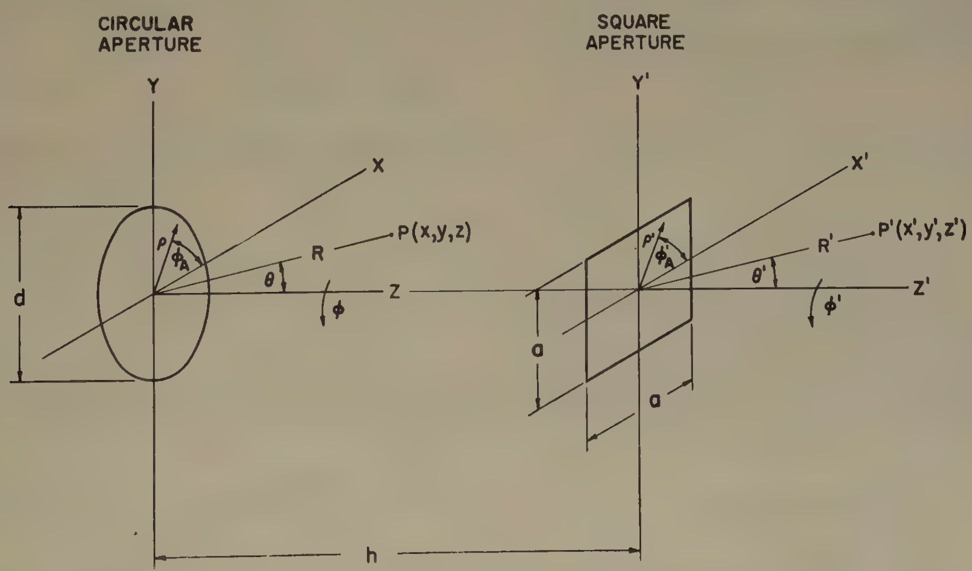

Fig. 1—Equivalent system.

Circular Aperture Radiation Patterns

A general integral for the field intensity in Fraunhofer region of a uniform phase-front aperture is given by

$$\frac{j}{\lambda R} e^{-j\frac{2\pi R}{\lambda}} \int_A F(x, y) e^{j\frac{2\pi}{\lambda}\sin\theta(x\cos\phi+y\sin\phi)} dxdy.$$

This is in the form used by Silver,[1] where changes in the obliquity factor are neglected due to the highly directive nature of the apertures under consideration. From this general integral, Silver derives a general expression for the aperture diffractivity in the Fraunhofer region of a circular aperture with a field-intensity distribution of the form $(1-\rho_n^2)^p$. This general expression[2] is

$$S(v) = \frac{\pi d^2}{2} \int_0^1 (1 - \rho_n^2)^p \rho_n J_0(v\rho_n) d\rho_n$$

* Decimal classification: R325.7. Original manuscript received by the IRE, December 7, 1953; revised manuscript received, March 9, 1954.

† Radio Research Div., The Western Union Telegraph Co., New York, N. Y.

[1] S. Silver, "Microwave Antenna Theory and Design," McGraw-Hill Book Co., Inc., New York, N. Y., p. 173; 1949.

[2] *Ibid.*, p. 194.

$$= \frac{\pi d^2}{4(p+1)} \Lambda_{p+1}(v),$$

where $v = (\pi d/\lambda) \sin \theta$. The normalized radiation pattern of a circular aperture is thus $\Lambda_{p+1}(v)$. Spherical Bessel functions of the form $\Lambda_p(v)$ are tabulated.

A uniformly illuminated circular aperture ($p=0$) then has a normalized radiation pattern given by $\Lambda_1(v)$ and a normalized power radiation pattern given by $\Lambda_1^2(v)$. This is a reference condition to be considered later.

The illumination of the circular aperture defined by the parabolic antenna in an actual system is tapered. This taper is assumed to be of the form $(1-\rho_n^2)^{1/2}$. The normalized radiation pattern is then given by $\Lambda_{3/2}(v)$ and the normalized power radiation pattern by $\Lambda_{3/2}^2(v)$.

Spillover Power

Only that portion of the power radiation pattern of the circular aperture that is intercepted by the square aperture represents useful power in the system. An expression is derived which represents the efficiency of the parabolic antenna and the loss due to spillover between the circular aperture and the square aperture.

When the radiation intensity in watts per solid angle is given by $P(\theta, \phi)$, then the total power radiated is

$$P_t = \int_0^{2\pi} \int_0^{\pi} P(\theta, \phi) \sin \theta d\theta d\phi.$$

For a circularly symmetrical radiation pattern, the integral reduces to

$$P_t = 2\pi \int_0^{\pi} P(\theta) \sin \theta d\theta.$$

$P(\theta)$ can be expressed as $KS^2(\theta)$. The upper limit of the integration can be reduced to some value of Θ which does not violate the approximation made in deriving $S(\theta)$ and still includes substantially all of the power in the radiation pattern.

$$P_t = 2\pi \int_0^{\Theta} KS^2(\theta) \sin \theta d\theta$$

Changing the variable from θ to a reduced angle $v = (\pi d/\lambda) \sin \theta$, the integral becomes

$$P_t = \frac{K\lambda^2}{\pi^2 d^2} \int_0^V S^2(v) 2\pi v dv,$$

where $V = (\pi d/\lambda) \sin \Theta$. When $S^2(v)$ is normalized to a peak value of one and the constants are combined

$$P_t = K' \int_0^V S_n^2(v) 2\pi v dv.$$

This expression shows that the power radiated is proportional to the volume enclosed by the power radiation pattern.

Consider then the normalized $\Lambda_1^2(v)$ radiation pattern of a uniformly illuminated circular aperture. The total power enclosed by this pattern is

$$P_t = K'\{4\pi[1 - J_0^2(V) - J_1^2(V)]\}.$$

The values of d, λ, and Θ give a value of V large enough so that in the limit P_t approaches $4\pi K'$. This is a reference condition where there are no losses associated with the aperture; that is, the power radiated is equal to power supplied.

If the same total power ($4\pi K'$) is supplied to the actual parabolic antenna, there are losses involved. It is customary, in considering conventional parabolic antennas, to assume the peak radiation intensity to be reduced to 0.65. This figure represents all of the losses in the parabolic antenna, including a reduction due to the illumination taper. The power radiated is then represented by a circularly symmetrical 0.65 $\Lambda_{3/2}^2(v)$ pattern.

The power intercepted by the square aperture is obtained by integrating that portion of the 0.65 $\Lambda_{3/2}^2(v)$ pattern that is intercepted. When that portion of the pattern is approximated by 0.65 $\{1-(1-\delta)[(x_n')^2+(y_n')^2]\}^2$, where δ is the relative intensity at the edge of the aperture in a principal plane, and the solid angle is approximated by $(a^2/4h^2)dx_n'dy_n'$; the power intercepted is

$$P_i = \frac{0.65K'\pi^2a^2d^2}{\lambda^2h^2} \cdot \int_0^1 \int_0^1 \{1 - (1 - \delta)[(x_n')^2 + (y_n')^2]\}^2 dx_n'dy_n'.$$

Integrating gives

$$P_i = \frac{0.65K'\pi^2a^2d^2}{\lambda^2h^2} [1 - 4/3(1 - \delta) + 28/45(1 - \delta)^2].$$

Thus the ratio of the power intercepted by the square aperture to the total power supplied to the parabolic antenna is

$$\frac{P_i}{P_t} = \frac{0.65\pi a^2d^2}{4\lambda^2h^2} [1 - 4/3(1 - \delta) + 28/45(1 - \delta)^2].$$

Fraunhofer Gain of the Square Aperture

The Fraunhofer gain of the nonuniformly illuminated square aperture is given by $G = G_u\mathcal{G}$. $G_u = 4\pi a^2/\lambda^2$ is the gain of the aperture uniformly illuminated, and $\mathcal{G} = S_2^2(0, 0)P_1/S_1^2(0, 0)P_2$ is the gain factor. $S_1(0, 0)$ and $S_2(0, 0)$ are the maximum values of the aperture diffractivities, at θ' and ϕ' equal to zero, for uniform and nonuniform illuminations respectively. P_1 and P_2 are the powers supplied to the square aperture with uniform and nonuniform illuminations, for the same peak value of illumination in the aperture.

For a square aperture with a $\{1-(1-\delta)[(x_n')^2 +(y_n')^2]\}$ intensity taper, the aperture diffractivity is given by

$$S(\theta', \phi') = \frac{a^2}{4}\int_{-1}^{+1}\int_{-1}^{+1}\{1-(1-\delta)[(x_n')^2+(y_n')^2]\}$$
$$\cdot e^{jux_n'\cos\phi'}e^{juy_n'\sin\phi'}dx_n'dy_n',$$

where $u=(\pi a/\lambda)\sin\theta'$. Integrating and substituting the limits yields

$$S(\theta', \phi') = a^2(2\delta-1)\left(\frac{\sin A}{A}\right)\left(\frac{\sin B}{B}\right) + \frac{2a^2}{3}(1-\delta)\left[\left(\frac{\sin B}{B}\right)\Lambda_{3/2}(A) + \left(\frac{\sin A}{A}\right)\Lambda_{3/2}(B)\right],$$

where $A=u\cos\phi'$ and $B=u\sin\phi'$. Setting θ' and ϕ' equal to zero gives the maximum value of the aperture diffractivity $S(0, 0)=a^2(2\delta+1)/3$. Then for uniform illumination, when $\delta=1$, $S_1(0, 0)=a^2$. For nonuniform illumination, $S_2(0, 0)=a^2(2\delta+1)/3$.

From the expression for the power intercepted by the square aperture, it follows that

$$\frac{P_1}{P_2} = \frac{1}{[1-4/3(1-\delta)+28/45(1-\delta)^2]}.$$

Substituting the values of the diffractivities and powers into the expression for the gain factor gives

$$\mathcal{G} = \left(\frac{2\delta+1}{3}\right)^2\left[\frac{1}{1-4/3(1-\delta)+28/45(1-\delta)^2}\right].$$

The Fraunhofer gain of the square aperture under consideration is then given by

$$G = \left(\frac{4\pi a^2}{\lambda^2}\right)\left(\frac{2\delta+1}{3}\right)^2 \cdot \left[\frac{1}{1-4/3(1-\delta)+28/45(1-\delta)^2}\right].$$

The normalized power radiation pattern, in any plane, is obtained by normalizing $S(\theta', \phi')$ to a peak value of one and squaring to give

$$\left\{\left(\frac{6\delta-3}{2\delta+1}\right)\left(\frac{\sin A}{A}\right)\left(\frac{\sin B}{B}\right) + \left(\frac{2-2\delta}{2\delta+1}\right) \cdot \left[\left(\frac{\sin A}{A}\right)\Lambda_{3/2}(B) + \left(\frac{\sin B}{B}\right)\Lambda_{3/2}(A)\right]\right\}^2.$$

Setting $\phi'=0$, the normalized power radiation pattern in a principal plane becomes

$$\left[\left(\frac{4\delta-1}{2\delta+1}\right)\left(\frac{\sin u}{u}\right) + \left(\frac{2-2\delta}{2\delta+1}\right)\Lambda_{3/2}(u)\right]^2.$$

Another Point of View

Up to this point, the discussion has been concerned with the radiation pattern of the nonuniformly illuminated circular aperture, the spillover power between apertures, and the Fraunhofer gain of the nonuniformly illuminated square aperture. These factors have been related to one another in the expressions derived in the preceding sections. The product of P_i/P_t and G represents the Fraunhofer gain of the system over the power supplied to the system and is given by

$$\left\{\left(\frac{0.65\pi a^2 d^2}{4\lambda^2 h^2}\right)[1-4/3(1-\delta)+28/45(1-\delta)^2]\right\} \cdot \left\{\frac{\left(\frac{4\pi a^2}{\lambda^2}\right)\left(\frac{2\delta+1}{3}\right)^2}{1-4/3(1-\delta)+28/45(1-\delta)^2}\right\}.$$

If the terms in this expression are combined and rearranged, the expression reduces to

$$\left[0.65\left(\frac{\pi^2 d^2}{\lambda^2}\right)\right]\left[\frac{\lambda^2}{16\pi^2 h^2}\right]\left[\frac{4\pi a^2}{\lambda^2}\right]\left[\frac{4\pi a^2}{\lambda^2}\left(\frac{2\delta+1}{3}\right)^2\right].$$

The first term is the Fraunhofer gain of a parabolic antenna, designated as G_1.

The second term is the free-space attenuation between isotropic radiators separated by a distance h, designated α_h.

The third term is the Fraunhofer gain of a uniformly illuminated square aperture, designated as G_2.

The fourth term is the Fraunhofer gain of a nonuniformly illuminated square aperture, designated G_3. The nonuniform illumination arises from the parabolic antenna radiation pattern.

Fresnel Correction Factors

Normally antennas operate in the Fraunhofer region and the gain in this region is constant and independent of distance. In the passive reflector antenna system, however, component antennas operate near the transition between Fresnel and Fraunhofer regions, and on occasion, within the Fresnel region. The gain in the Fresnel region is not constant but a function of distance. Thus a correction factor must be developed to modify the Fraunhofer gain when the antenna operates in the Fresnel region. The correction factor is taken as

$$\frac{|S_{\max}|^2_{\text{Fresnel}}}{|S_{\max}|^2_{\text{Fraunhofer}}}.$$

Two apertures in the system under consideration operate over the distance h, which may be within their Fresnel regions. Thus correction factors for the Fraunhofer gains G_1 and G_2 must be developed. It is not necessary to develop a correction factor for G_3 since this aperture operates over a distance which is assumed to be well into the Fraunhofer region.

The correction factor for the circular aperture illuminated with a $(1-\rho_n^2)^{1/2}$ taper is developed in the following manner. The maximum value of the diffractivity of a circular aperture in the Fresnel region is given by Silver[3] as

[3] *Ibid.*, p. 198.

$$S(z) = \int_0^{2\pi} \int_0^{d/2} F(\rho, \phi_A) e^{-j\frac{\pi\rho^2}{\lambda z}} \rho d\rho d\phi_A.$$

Substituting $F(\rho, \phi_A) = (1 - 4\rho^2/d^2)^{1/2}$, the illumination taper since $\rho_n = 2\rho/d$; and integrating yields

$$S(z) = \frac{\pi d^2}{2} e^{-jm} \left\{ \sum_{n=0}^{n=\infty} \frac{(-1)^n m^{2n}}{(4n+3)[(2n)!]} + j \sum_{n=0}^{n=\infty} \frac{(-1)^n m^{2n+1}}{(4n+5)[(2n+1)!]} \right\},$$

where $m = \pi d^2/4\lambda z$. If the least distance h to be considered is taken as $\frac{1}{4}$ of $2d^2/\lambda$; then with $z = h$, the maximum value that m can have is 1.57. In practice, the values of m are less than 1.57 and the two series converge rapidly in this region. Thus the absolute value of the diffractivity can be approximated by

$$\left| S_{\max} \right|_{\text{Fresnel}} = \left(\frac{\pi d^2}{2} \right) \left\{ \frac{1}{9} \left[1 - 0.0684 \left(\frac{\pi d^2}{4\lambda h} \right)^2 \right] \right\}^{1/2}.$$

The Fraunhofer diffractivity of this same aperture has been shown to be

$$S(v) = \left(\frac{\pi d^2}{4} \right) \left(\frac{2}{3} \right) \Lambda_{3/2}(v).$$

The maximum value occurs at $v = 0$ and is

$$\left| S_{\max} \right|_{\text{Fraunhofer}} = \left(\frac{\pi d^2}{4} \right) \left(\frac{2}{3} \right).$$

The correction factor then becomes

$$\left[1 - 0.0684 \left(\frac{\pi d^2}{4\lambda h} \right)^2 \right].$$

The correction factor for the uniformly illuminated square aperture is developed in a similar manner and is given by

$$\left[\frac{C^2\left(\frac{a}{\sqrt{2\lambda h}} \right) + S^2\left(\frac{a}{\sqrt{2\lambda h}} \right)}{\frac{a^2}{2\lambda h}} \right]^2,$$

where C and S are cosine and sine Fresnel integrals of the form

$$C(x) = \int_0^x \cos\left(\frac{\pi}{2} t^2 \right) dt.$$

Reflector Efficiency

There is a reduction in the maximum possible gain from a plane reflector by other factors which have not been considered. One of these factors is the heat loss caused by currents on the surface of the reflector. This loss is estimated to be of the order of one per cent, or η heat loss $= 0.99$.

Another factor is the reduction in effective area of the aperture due to the discontinuity in the field near the edge of the reflector. It is estimated that the area within a quarter wavelength of the edge is only 50 per cent efficient. The efficiency due to the edge effect is η edge effect $= (1 - \lambda/2a)^2$.

These efficiency factors are arbitrary and have been estimated on the conservative side. Other factors which effect the efficiency of the reflector are not considered because the magnitude of their effects is negligible.

System Gain

Earlier an expression including the Fraunhofer gains of the apertures and the free-space attenuation between apertures was derived. If the Fraunhofer gains are modified by the Fresnel correction factors and the reflector efficiency is included, this expression gives the system gain of a passive reflector antenna system.

$$G = G_1 \alpha_h G_2 G_3 \eta$$

where

$$G_1 = \left[0.65 \left(\frac{\pi^2 d^2}{\lambda^2} \right) \right] \left[1 - 0.0684 \left(\frac{\pi d^2}{4\lambda h} \right)^2 \right],$$

$$\alpha_h = \frac{\lambda^2}{16\pi^2 h^2},$$

$$G_2 = \left(\frac{4\pi a^2}{\lambda^2} \right) \left[\frac{C^2\left(\frac{a}{\sqrt{2\lambda h}} \right) + S^2\left(\frac{a}{\sqrt{2\lambda h}} \right)}{\frac{a^2}{2\lambda h}} \right]^2,$$

$$G_3 = \left(\frac{4\pi a^2}{\lambda^2} \right) \left(\frac{2\delta + 1}{3} \right)^2,$$

and

$$\eta = (0.99)(1 - \lambda/2a)^2.$$

This expression can be reduced, for easier calculation in decibels, to

$$G = 20 \log (d/\lambda)(2\delta + 1)[C^2(x) + S^2(x)] \cdot (1 - \lambda/2a)(1 - 0.0684m^2)^{1/2} + 4.50$$

where $x = a/\sqrt{2\lambda h}$, $m = \pi d^2/4\lambda h$, and $\delta = \Lambda_{3/2}(\pi da/2\lambda h)$.

It is interesting to compare the gain of this system with that of a system using a reflector defining a circular aperture. If, in the systems to be compared, the diameter equals a; then as h increases and δ approaches one, the ratio of the system gains approaches the ratio of the reflector areas squared. This ratio is $16/\pi^2$. Thus the "corners" of a reflector can contribute as much as 2.1 decibels of system gain.

The system gains at 4,000 mc are plotted Figs. 2, 3, and 4 for passive reflector antenna systems with square apertures fed by 4-, 6-, and 8-foot parabolic antennas respectively. These curves show that as the distance of separation decreases, the gain increases until the maximum is reached. This increase in gain occurs because the reflector intercepts more power as the distance de-

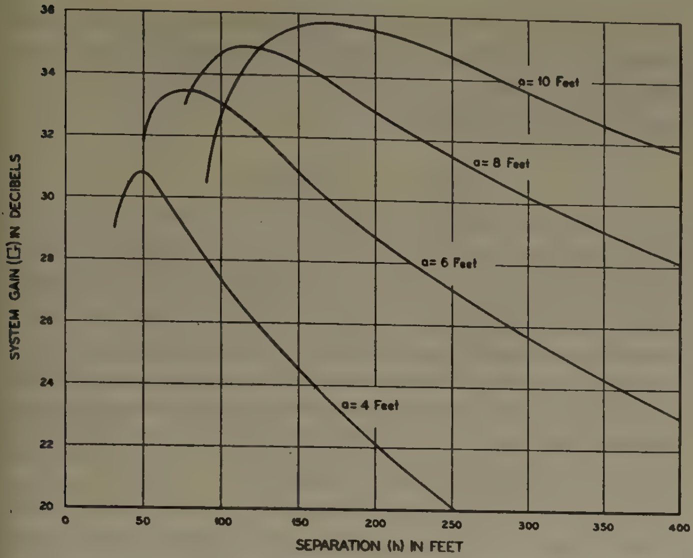

Fig. 2—System gain of a passive reflector antenna system fed by a 4-foot parabolic antenna at 4,000 mc.

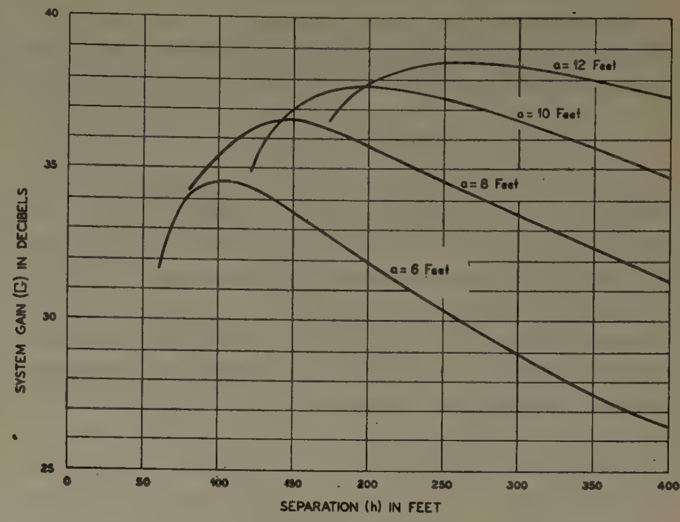

Fig. 3—System gain of a passive reflector antenna system fed by a 6-foot parabolic antenna at 4,000 mc.

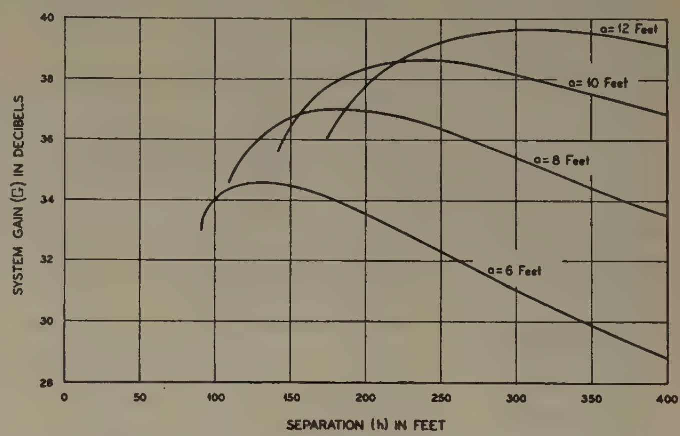

Fig. 4—System gain of a passive reflector antenna system fed by a 8-foot parabolic antenna at 4,000 mc.

creases and this increasing power overshadows a decreasing gain factor. However, even though more power is intercepted as the distance decreases beyond the point where the maximum occurs, the gain in this region decreases. This occurs because the reflector has exceeded the size of a modified first Fresnel zone; that is, all of the energy in the reflector aperture does not contribute in phase at a distant point. In this region of decreasing gain, the reflector is larger than the optimum size and an increase in gain can be realized by decreasing the size of the reflector. Another method of increasing the gain in this region is to curve the surface of the reflector so that the system approaches an offset-fed parabolic antenna—on a grand scale.

Since the region to the left of the maximum does not represent an operating condition, there being little justification for using a reflector larger than the optimum size; the accuracy of the curves in this region is of little concern. It is estimated that the accuracy of these curves in the regions representing practical operating conditions is of the order of plus or minus 0.5 decibel.

Experimental Gains

The gains of four passive reflector antenna systems were compared with the gain of a conventional parabolic antenna mounted on the tower adjacent to the plane reflector. The results of these tests are summarized in Table I, below.

The average difference between experimental and theoretical values is less than 0.5 decibel. Two experimental gains are higher and two experimental gains are lower than the theoretical values. The maximum difference is 0.89 decibel.

TABLE I

Comparison of Experimental and Theoretical Gains of Passive Reflector Antenna Systems Tested at Monsey Heights, N. Y.

Diameter of Paraboloid	Distance of Separation	Size of Reflector Aperture	Frequency	System Gain Over an Isotropic Antenna	System Gain Over a 4-foot Parabolic Antenna	
				(Theoretical)	(Theoretical)	(Experimental)
d feet	h feet	a feet	mc	G db	G-32.43 db	db
6	90.25	6	4,060	34.45	+2.02	+2.50
6	140.80	6	4,060	33.87	+1.44	+0.55
4	90.25	6	4,060	33.40	+0.97	+0.70
4	140.80	6	4,060	31.39	−1.04	−0.75

The magnitude of these differences is not unreasonable when the variables in the gain analysis and in the experimental determination of the gain are considered.

Side-Lobe Levels

The side-lobe levels in a passive reflector antenna system vary depending upon the illumination at the edge of the reflector. This illumination is a function of the component antennas and their separation.

The radiation pattern of a passive reflector antenna system, where $\delta = 0.882$, was measured at Monsey. This pattern is plotted in Fig. 5 with the theoretical pattern included for comparison.

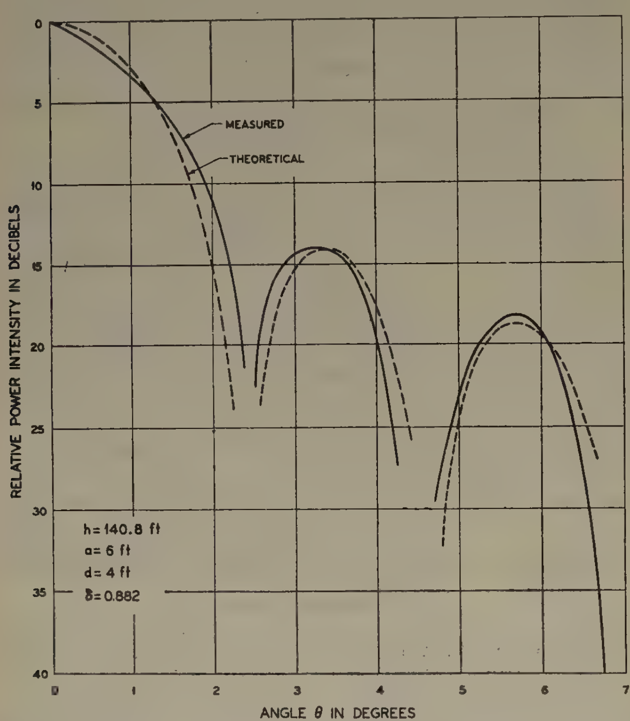

Fig. 5—Comparison of the theoretical and experimental principal plane radiation patterns of a passive reflector antenna system at 4,060 mc.

If side-lobe levels are to be of major concern, there are several methods that can be used to reduce them. At the expense of system gain, the "corners" of a square reflector can be eliminated. Since a uniformly illuminated square aperture has a first side-lobel level of 13.2 db while a uniformly illuminated circular aperture has a first side-lobe level of 17.6 db, this measure is particularly effective where the illumination is close to uniform.

Another method is to decrease the level of the illumination at the edge of the reflector relative to that at the center. This can be done by increasing the size of the plane reflector, the parabolic antenna, or both. Under most normal operating conditions this will also increase the system gain. The exceptions are, of course, when the systems are operating at or to the left of the maxima on the gain curves in Figs. 2, 3, and 4.

Mutual Coupling

The mutual coupling among passive reflector antenna systems is relatively complex. Coupling exists, to some degree, among all of the components in the system. However, the open nature of the system makes it particularly susceptible to coupling by reflections from local terrain, nearby objects, and the tower itself. The presence of portions of the tower structure in the path of radiation within the antenna system can contribute an appreciable amount of coupling.

If reasonable care is exercised in selecting a site, and a tower construction is employed which avoids large reflecting surfaces; local coupling due to reflections can be minimized to the point where the coupling between the parabolic antennas alone is a dominant factor. Measurements show that local coupling between passive reflector antenna systems of the order of 50 to 60 db can be obtained.

Coupling from distant sources, however, is a far more serious problem because of the probability that the desired signal will fade while the undesired signal remains at normal strength. Measurements during time of normal propagation indicated that coupling of the order of 40 db could be expected.

Although this figure could be improved by shielding between systems, it is doubtful if it could be improved to a point where a satisfactory carrier-to-noise ratio could be maintained during periods of fading if the same frequency were to be employed more than once at a single repeater installation.

Conclusion

This paper has discussed principally the gains of passive reflector antenna systems. The system gain has been derived in general terms and curves similar to those included for operation at 4,000 mc can be calculated for systems operating at other frequencies in the microwave region.

In the process of deriving the system gain, the normalized power radiation pattern was found. From this pattern other radiation characteristics such as beamwidths, positions of nulls, and the positions and values of side-lobe peaks can readily be determined.

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The Theory and Application of the Radiation Mutual-Coupling Factor*

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Summary—**The network equations for determining mutual impedance between a pair of radiating elements are analyzed and a new equation is determined in which the mutual impedance is expressed in terms of the self-impedances of the radiating elements and a new parameter designated as the radiation mutual-coupling factor M_r. A technique is outlined whereby this new parameter can be isolated and accurately measured despite the presence of discontinuities at the base of the radiating element to which a slotted line is connected. The technique is demonstrated for the case of a pair of identical quarter-wave monopoles and the results compared with the theroetical values interpolated from King's approximate second-order solution. The technique is then applied to the case of a pair of identical axial mode helical radiators.**

The Network Equations and the Radiation Mutual-Coupling Factor

CONSIDER an array of radiators as a network of intercoupled elements, then an expression for the terminal impedance can be derived from circuit theory. As an example, the case of a driven radiator over a ground plane and a similar parasitic element shown in Fig. 1 will have voltage and current expressions as follows:[1,2]

$$\left.\begin{aligned} V_1 &= I_1Z_{11} + I_2Z_{22} \\ 0 &= I_1Z_{21} + I_2Z_{22} \end{aligned}\right\} \quad (1)$$

V_1 = Terminal voltage of radiator #1
I_1 = Terminal current of radiator #1
I_2 = Terminal current of radiator #2
Z_{11} = Self-impedance of radiator #1
Z_{22} = Self-impedance of radiator #2
$Z_{12} = Z_{21}$ (assumed) = Mutual impedance between radiator #2 and radiator #1
$Z_t = V_1/I_1$, Terminal impedance of radiator #1.

In solving (1) above, and assuming that $Z_{12} = Z_{21} = Z_m$, the following expression results:

$$Z_m^2 = (Z_{11} - Z_t)Z_{22}. \quad (2)$$

As a first approximation Z_{11} may be determined by open-circuiting radiator #2 and assuming that I_2 will become zero. There is an inaccuracy involved this in assumption; for, open-circuiting the conductor at the base of the radiating element will not cause the current I_2 to become zero since a capacitive reactance will be created which is a function of the gap width and conductor diameter. The displacement current between other portions of the antenna is not included since I_2 refers only to the current at the radiator terminals. Most theoretical derivations of self- or mutual-impedance do not take this into account since they are for isolated radiators minus all discontinuities at the terminals. Actually, the proximity of an open-circuit parasitic element will distort the current distribution causing a change in the self-impedance[3] of the driven radiator and the amount of distortion varies as the separation distance between a driven and a parasitic element.

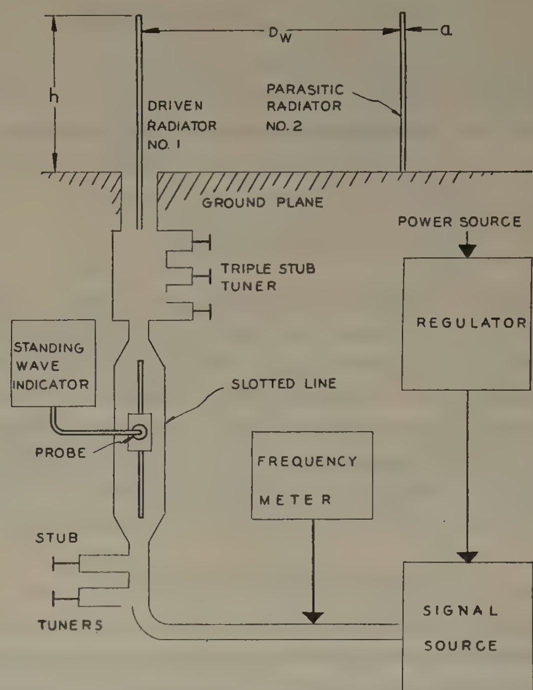

Fig. 1—Schematic of measurements apparatus for determining mutual impedance or radiation mutual-coupling factor.

Therefore, the network equations may be more clearly stated if (2) is rewritten as follows:

$$Z_m^2 = (Z_{1oc2} - Z_{1sc2})Z_{2oc1} \quad (3)$$

where,

Z_{1oc2} = Self-impedance of radiator #1 in the presence of radiator #2 open-circuited
Z_{2oc1} = Self-impedance of radiator #2 in the presence of radiator #1 open-circuited
Z_{1sc2} = Terminal impedance of radiator #1 in the presence of radiator #2 short-circuited.

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[1] J. D. Kraus, "Antennas," McGraw-Hill Book Co., Inc., New York, N. Y., chaps. 7 and 10; 1950.

[2] R. King, H. R. Mimno, and A. H. Wing, "Transmission Lines, Antennas, and Waveguides," McGraw-Hill Book Co., Inc., New York, N. Y.; 1945.

[3] Impedance with no other radiator present, or removed to such a distance as to have negligible effect.

The representative coupling network between the two radiating elements may actually be very complex. No matter how complex it may be, it can according to theory be reduced to a four-terminal network. However, as a mathematical convenience, a complex number can be represented as a product of two other complex numbers as follows:

$$Z_{1oc2} = Z_{11}T_{1oc2}$$

$$Z_{2oc1} = Z_{22}T_{2oc1}$$

$$Z_{1sc2} = Z_{11}T_{1sc2}.$$

T_{1oc2}, T_{2oc1}, and T_{1sc2} are complex numbers that modify the self-impedance of the radiators because of the presence of another radiator in its vicinity. These expressions hold true for a fixed geometry of the radiators at a single frequency. Rewriting (3):

$$Z_m^2 = (Z_{11}T_{1oc2} - Z_{11}T_{1sc2})Z_{22}T_{2oc1}$$

$$Z_m = [(T_{1oc2} - T_{1sc2})T_{2oc1}]^{1/2}\sqrt{Z_{11}Z_{22}}.$$

Notice that the term in the bracket merely contains transformation complex numbers which deal solely with the coupling interactions between the radiating elements; and it will be designated with the symbol M_r. Hence,

$$Z_m = M_r\sqrt{Z_{11}Z_{22}}. \tag{4}$$

In (4), M_r is defined as the *Radiation Mutual Coupling Factor* and is a dimensionless complex number. This quantity has its counterpart in the coupling between coils and is well known as the coefficient of coupling "k," which is a pure dimensionless number. The radiation mutual-coupling factor[4] tells us specifically what the coupling interaction actually is between two radiating elements and represents a true figure of merit.

The Measurement Technique

In actual practice, means must be provided to support radiating elements as well as connecting a transmission line or measurement instrument to them. Supports and connectors are sources of discontinuities which in the case of mutual- or self-impedance measurements cause discrepancies between measured and theoretical results.[5]

If a typical mutual-impedance-measurement arrangement is considered as shown in Fig. 1 (with the exception of the triple-stub tuner), much can be done to minimize and compensate for terminal discontinuities. As for the parasitic element, no problem exists for the short-circuit condition. However, as previously indicated, the open-circuit condition does present a disturbing capacitive quantity at the terminus and this must be reduced to as small a value allowable by the physical geometry of the radiator proper. Directing our attention to the driven element, it is almost impossible to eliminate all discontinuities between the driven element and the slotted measuring line. Let us assume that discontinuities introduced by the radiator-base support and connector are present and that instead of measuring the true self-impedance of an isolated element, one measures some other quantity of impedance which shall be designated as Z_{11}'. Again, the self-impedance of the radiator can be represented as the product of two complex quantities, such as:

$$Z_{11} = Z_{11}'T_{d1}.$$

Furthermore, another complex quantity can be selected such that:

$$(Z_{11}'T_{d1})T_{n1} = R_0 \tag{5}$$

where R_0 is the characteristic impedance of the slotted line. At 1,500 mc T_{n1} could be represented by a triple-stub tuner and referring again to Fig. 1, it is merely inserted between the radiating element and the measuring slotted line. The triple-stub tuner is adjusted to a value of voltage-standing-wave ratio of as near unity as possible and this is done with the driven element isolated from any other element. Similarly, the same can be done with radiator #2 such that:

$$Z_{22}'T_{d2}T_{n2} = R_0. \tag{6}$$

It must be realized that when the above adjustment is made with the triple-stub tuner, radiator reactances, differences between radiator resistance and slotted-line-characteristics impedance, as well as discontinuities at the radiator-base terminus are being matched. Subsequent measurements must therefore be referred to the load end of the slotted line instead of the base of the radiator. With this initial adjustment, the parasitic element is introduced in the vicinity of the driven element producing coupling interaction which causes the initial matched condition to be disturbed. However, the mutual impedance cannot be measured since the triple-stub tuner has been inserted between the driven element and the measuring line. Whatever this quantity may be, let us designate it as Z_m'. The result of this measurement is:

$$(Z_m')^2 = (Z_{11}'T_{d1}T_{1oc2}T_{n1} - Z_{11}'T_{d1}T_{1sc2}T_{n1})Z_{22}'T_{d2}T_{2oc1}T_{n2}.$$

But originally the triple-stub tuner was adjusted for conditions described in (5) and (6) and after proper substitution and simplification, the following results:

$$Z_m' = [(T_{1oc2} - T_{1sc2})T_{2oc1}]^{1/2}R_0$$

or,

$$M_r = Z_m'/R_0. \tag{7}$$

This last expression states that the normalized value of the quantity Z_m' which is measured after having initially

[4] R. King, "Coupled antennas and transmission lines," Proc. I.R.E., vol. 31, pp. 626–639; November, 1943. At the completion of the work, it came to the writer's attention that here King had also made reference to a similar dimensionless quantity.

[5] D. D. King, "The measured impedance of cylindrical dipoles," *Jour. Appl. Phys.*, vol. 17; October, 1946.

adjusted the triple-stub-tuner network for a vswr of unity when the driven element was isolated is identical with the radiation mutual-coupling factor M_r.

The Radiation Mutual-Coupling Factor Between Monopoles

If two monopoles are constructed with good machining tolerances and then properly fitted with necessary supports and connectors, it would be found in making mutual impedance measurements that the results will not agree with theoretical values primarily because of the presence of discontinuities at the terminal zone. The first step in attempting to eliminate such terminal discontinuities is to minimize those appearing at the base of the parasitic element. So the connector was removed from the base and the radiator was supported either by soldering it directly to the ground plane (short-circuit condition) or by insertion of a good piece of dielectric between the base and the ground plane (open-circuit condition) assuring that the capacity introduced was not too large.

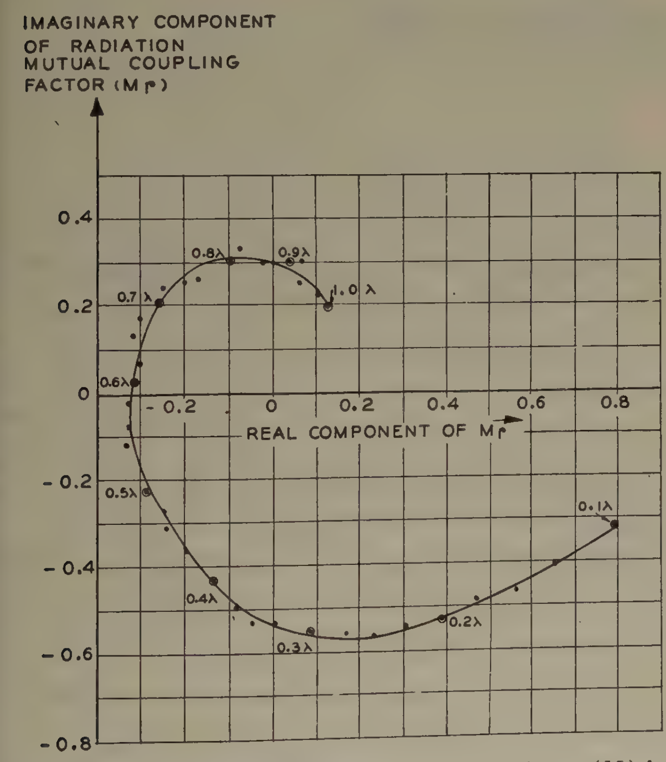

Fig. 2—*R-I* diagram of radiation mutual-coupling factor (M_r) between a pair of equal monopoles
$L/D=61.5$ ($\Omega=11.0$)
$Z_{11}=Z_{22}$ ·—· measured values every 0.025λ.
$f=1{,}500$ mc

The procedure as previously outlined was then used in determining the radiation-coupling factor and the results are shown in Fig. 2. These results were then multiplied by the self-impedance of one of the elements since in this case both the driven and the parasitic elements were identical. King's[6] approximate second-order solution for a radiator type $\Omega=11.0$, where $Z_{11}=42.68-j21.02$ was used as the self-impedance value. Note that the theoretical value determined by King was divided by two for the case of the monopoles. The resulting mutual impedance values were then plotted as shown by Fig. 3.

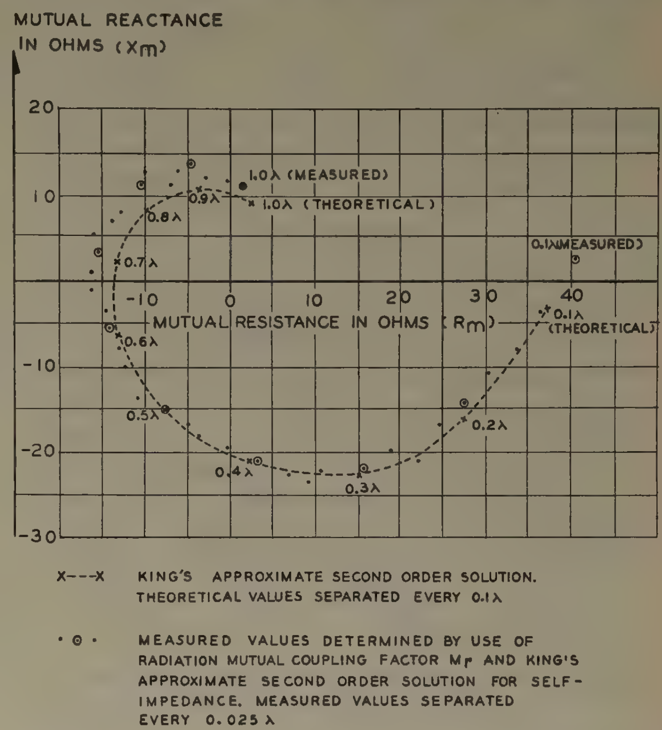

Fig. 3—*R-X* diagram of the mutual impedance for a pair of identical λ/4 radiators
$L/D=61.5$ ($\Omega=11.0$)
$Z_m=M_rZ_{11}$ where $Z_{11}=42.68+j21.02$ ($\Omega=11.0$).

Good agreement between measured values and King's approximate second-order theoretical solution values were obtained except for very small separation distances (approximately one tenth wavelength) and very large separation distances (in the order of one wavelength). The differences between measured and theoretical values for large separation distances can be attributed mostly to inaccuracies of measurement. At these larger separation distances the value of $|M_r|$ is less than 0.3 and the reliability of measurements at such levels is rather poor. However, for small separation distances there also exist differences between measured and theoretical values although the reliability of measurements is quite excellent. These differences may be attributed to the presence of small residual discontinuities, cumulative errors due to frequency instability, machining tolerances, the small base capacity present at the parasitic element, errors in the setting of the displacement distances or parallelism of the radiating elements. On the other hand, by examining Fig. 3, good agreement exists in the region between 0.3- and 0.6-wavelength separation. King's[6] report shows that for two identical radiators of half wavelength, irrespective of the L/D ratio, the values of resistive components of mutual impedance converge to a common value in the vicinity between 0.3- and 0.35-wavelength-separation distance. The reactance

[6] R. King, "Self and Mutual Impedances of Parallel Identical Antennas," Cruft Lab. Tech. Rep. No. 118; November, 1950.

components converge to a common value in the region between 0.5- and 0.55-wavelength separation. The absolute value of mutual impedance does not have any cross-over point like the resistive and reactance components but there does exist a convergence in the region

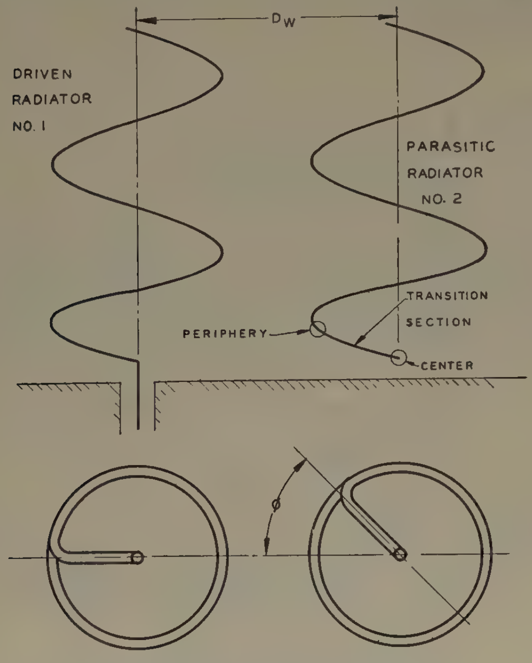

Fig. 4—Sketch of the helical radiator showing the transition section and the orientation angle ϕ.

of 0.55 wavelength. Similarly, the phase angle has a convergence between 0.4- and 0.45-wavelength-separation distance. Hence it could be concluded that the discrepancy present at the small separation distance could possibly be attributed to the fact that the theoretical second-order solution is approximate rather than exact. If the absolute value of mutual impedance or the phase angle of the mutual impedance are plotted against separation distance, it would be found that the curve deviates about a mean value which distinguishes itself from the case of the infinitely thin elements. These deviations were always noticed during the course of the measurements whether discontinuities were present at the terminal zone or not. King's theoretical work brings out these deviations which are consistent with the measured results presented. As a matter of interest, a radiation mutual-coupling-factor curve for two dissimilar elements, one a quarter-wave monopole and the other a third wavelength monopole, was determined but the results are not shown in this article. The difference in the measurement of the dissimilar monopoles in contrast with the two identical monopoles is that an extra set of measurements have to be effected in order to determine T_{2oc1}.

The Radiation Mutual-Coupling Factor and Helical Radiators

The helical radiator is a more complex type element than the linear monopole. In addition to the usual base discontinuities generally encountered, a portion of the first turn of the helix behaves very much like a transmission line due to its close proximity to the ground plane. See Fig. 4. Because of this short transition element, it becomes exceedingly difficult to measure and determine the inherent self-impedance of a helical radiator. At any rate, the radiation mutual-coupling factor, employing the technique previously described, was measured under various conditions. After studying and comparing the results of many measurements, it was decided to select the results measured under the following conditions:

(a) Elimination of the transition element between the feed point (center) and the periphery of the parasitic helix.
(b) Open- and short-circuiting the parasitic helix at the periphery.
(c) By use of the transition element, match the driven helix to its feed point by adjusting its displacement from the ground plane. Any residual discontinuity would be compensated by means of the triple stub tuner.

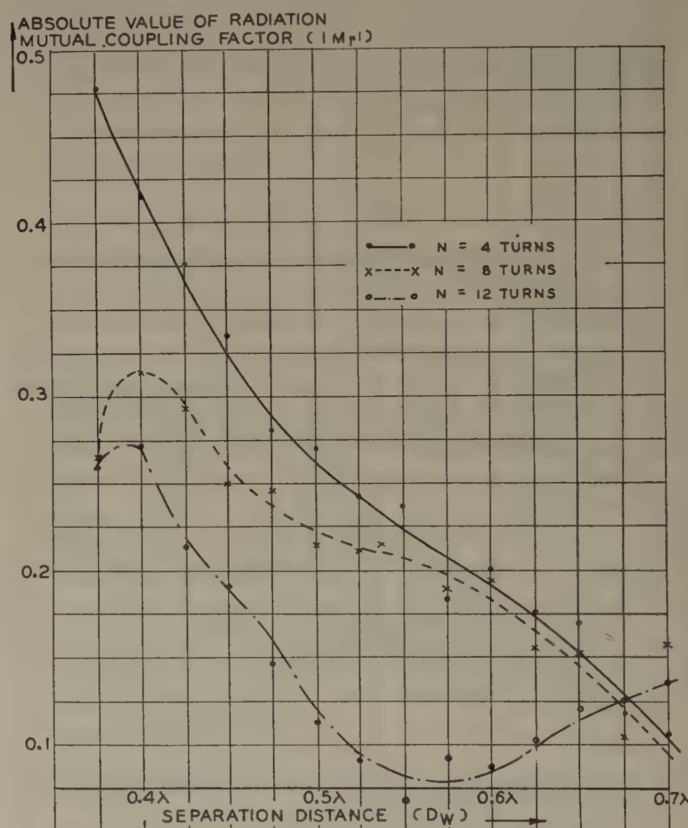

Fig. 5—Absolute value of radiation mutual-coupling factor ($|M_r|$) between a pair of axial mode helical radiators wound in the same sense plotted against separation distance (D_W)

$C_W = 1.0$ $\alpha = 12$ degrees
$f = 1{,}500$ mc $\phi = 0$ degrees.

The results of these measurements for helices of various turn lengths (number of turns) are shown on Fig. 5. As would be expected, the radiation mutual-coupling factor diminishes as the turn length is increased. This, of course, is brought about by the fact that the radiated energy becomes more concentrated along the helix axis

as the turn length is increased. In addition, observation should be made of the manner in which the value of $|M_r|$ oscillates about a mean value just as it occurred in the case of the monopoles with a finite radius.

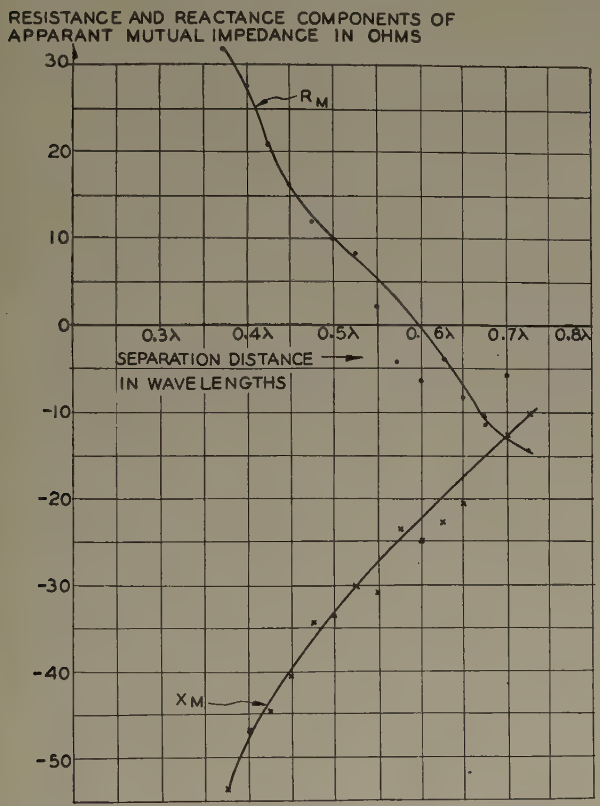

Fig. 6—R and X components of apparent mutual impedance plotted against separation distance for two identical 4-turn axial mode helical radiators wound in the same sense
$C_W = 1.0$ $\alpha = 12$ degrees $f = 1{,}500$ mc
$Z_m = M_r Z_{11}$ where $Z_{11} = 130.5 + j2.0$ (Apparent).

Fig. 6 shows an $R-X$ diagram of the apparent mutual impedance between a pair of identical 4-turn axial mode helical radiators. The term apparent is used since the inherent self-impedance of this type of radiator has neither been measured nor theoretically determined. Glasser and Kraus[7] determined the apparent self-impedance of a helix, similar to the ones used in determining their M_r, when connected to a 50-ohm termination and therefore their result was used in the calculations in (4). There are some discrepancies which should be mentioned in the above outlined measurements. The apparent self-impedance was measured at 425 mc where the ratio of conductor diameter to wavelength was 0.018. The radiation mutual-coupling factor was measured at 1,500 mc and the same ratio was approximately 0.016. In addition, the supporting dielectric structures were not identical.

Conclusions

Based on the results of the work outlined above, it is felt that the experimental and theoretical values of mutual impedances for the case of the monopoles are in reasonable agreement. Also, the radiation mutual-coupling factor as defined and measured is a true and accurate criterion for determining the coupling interaction between radiating elements. The technique outlined can be applied to most types of radiating elements provided proper measures are taken in minimizing discontinuities of the parasitic elements. It is also felt that the values of the radiation mutual-coupling factor for the case of axial mode helical radiators are considered adequate for design problems although there exists some question concerning the actual limit to which the transition element affects the measured values.

Acknowledgment

Sincere appreciation is directed towards J. D. Kraus for his generous guidance, interest, and encouragement as an advisor during the course of this work. Mention must also be made of the kind assistance rendered by K. E. Will in the design construction of the various radiating elements and ground planes used in the measurements.

[7] O. J. Glasser and J. D. Kraus, "Measured impedances of helical beam antennas," *Jour. Appl. Phys.*, vol. 19, p. 193; 1948.

Message Error in Diversity Frequency-Shift Reception*

G. FRANKLIN MONTGOMERY†, MEMBER, IRE

Summary—**Diversity methods are used to reduce receiving errors caused by carrier fading; for a keyed signal, such errors are easily specified in terms of the binary error. The attainable reduction in error depends upon the type of diversity selection used at the receiver. Several selection methods are examined, and the binary error is calculated for each, assuming narrow-band FSK transmission, fluctuation noise interference, and a Rayleigh distribution of carrier amplitudes. The results are used to determine the equivalent carrier-to-noise power gain for each diversity method. A numerical example is given in conclusion.**

Introduction

THE PURPOSE of this paper is to determine the reduction in error that can be achieved by using diversity reception of a narrow-band FSK transmission. It is assumed that n-receiver outputs are available, that the radio-frequency carrier amplitudes in these outputs are independent and follow a Rayleigh distribution,[1] and that the interference is band-limited white noise.[2,3]

Communication of a binary-coded signal may be analyzed as a sequence of time intervals equal in length to the shortest transmitted character. During each of these intervals, the transmitted frequency is one or the other of two possible frequencies. The binary error, p, is the average fraction of the total number of such intervals in the message during which the receiver makes an incorrect frequency determination.

We shall find it convenient to define the diversity power gain as

$$G_n(R) = \frac{R}{R_n}$$

where R is the average carrier-to-noise power ratio, and R_n is the average carrier-to-noise ratio that produces in the n-diversity system the same error as that produced by R in a single receiver. Thus, G_n is the reduction in transmitted carrier power that can be permitted if the n-diversity system is substituted for the single receiver.

To draw conclusions concerning diversity-receiver performance, it is necessary to consider selecting action of specific systems. Selection methods may be divided into two classes: those which examine various outputs at radio frequency, before detection, and those which examine demodulated, or signal, outputs.

Radio-Frequency Selection

It is assumed that the carrier amplitudes in the n outputs are independent. In general, the phases are independent also, and if no method of automatically adjusting the outputs to phase coherence exists, the best diversity system can do no more than select one of the outputs, rejecting the rest, on some amplitude or power basis. Three selection criteria are suggested. First, the system may select the output having the greatest carrier-to-noise ratio. Second, it may select the output having the greatest carrier amplitude (or the least noise). And third, it may select the greatest output amplitude, whether the output be carrier, noise, or a mixture of the two. To be most effective, the system should be able to make its selection in a period of time equal to or less than the interval of the shortest signal that will be transmitted. For the first two criteria, such selection speed is generally impossible. The determination of either carrier-to-noise ratio or carrier (or noise) alone implies that the system is capable of measuring carrier and noise separately in this short interval. If the system could perform this measurement with certainty, there would be no communication problem, and diversity reception would not be required. To the extent that the carrier-amplitude variation occupies a smaller bandwidth than the natural noise fluctuation, it is possible to measure carrier amplitude independently of noise and so reduce error using rf selection.[4]

Carrier-to-Noise Ratio Selection

For fading and noise conditions stated in the introduction, and for a narrow-band FSK system, it has been shown[5] that the binary error for a single receiver is

$$p = \frac{1}{2(1+R)},$$

* Decimal classification: R428. Original manuscript received by the IRE, July 30, 1953; revised manuscript received January 27, 1954.

† National Bureau of Standards, Washington, D. C.

[1] Lord Rayleigh, "The Theory of Sound," Dover Publications, New York, N. Y., pp. 35–42; 1945.

[2] S. Goldman, "Frequency Analysis, Modulation and Noise," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 327–330; 1948.

[3] V. D. Landon, "The distribution of amplitude with time in fluctuation noise," Proc. I.R.E., vol. 29, pp. 50–55; February, 1941. (See also V. D. Landon and K. A. Norton, "Discussion," Proc. I.R.E., vol. 30, pp. 425–429; September, 1942.)

[4] Suppose the system to operate with an rf bandwidth B, an rms carrier S_0, and an rms noise N_0. If the amplitude fluctuation of the carrier contains no component of frequency greater than $A/2$, where $A<B$, then it becomes possible to find (by filtering, say) a sum of carrier and noise, having the rms amplitude

$$\left(S_0^2 + \frac{A}{B}N_0^2\right)^{1/2} = S_0\left(1 + \frac{A}{BR}\right)^{1/2},$$

that constitutes a continuously more accurate measure of carrier amplitude as the ratio A/B approaches zero. The results found for the first two selection methods are the limits of error and diversity gain as A/B approaches zero.

[5] G. F. Montgomery, "A comparison of amplitude and angle modulation for narrow-band communication of binary-coded messages in fluctuation noise," Proc. I.R.E, vol. 42, pp. 447–454; February, 1954.

which is plotted in Fig. 1. This error is simply one-half the probability, q, that the demodulated output of the single receiver is not determined by the transmitted signal, and

$$q = \frac{1}{1+R}.$$

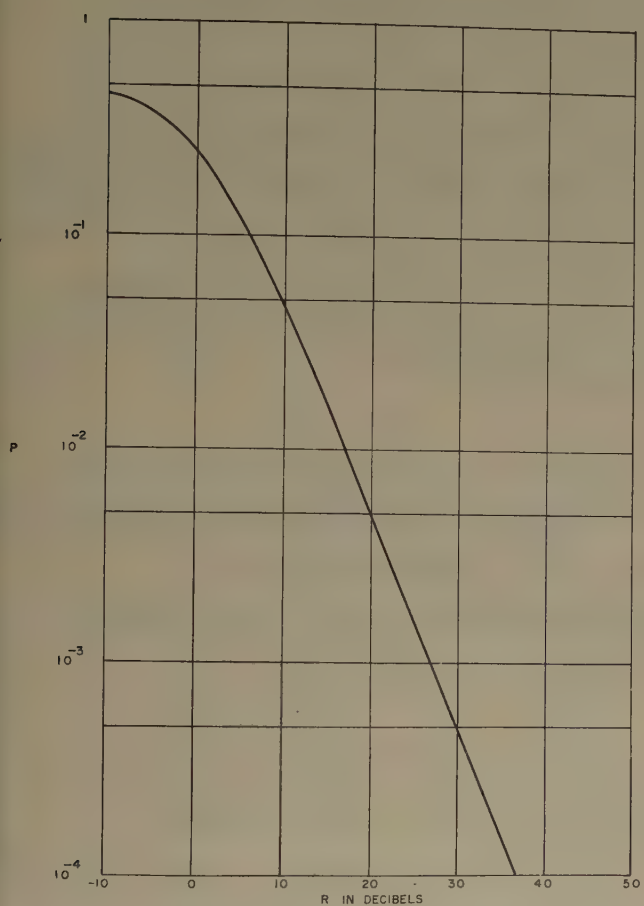

Fig. 1—Narrow-band FSK binary error vs average carrier-to-noise ratio, Rayleigh fading.

Since the carrier and noise amplitudes in the separate receivers of the n-diversity system are assumed independent, the probability that all n demodulated outputs are simultaneously indeterminate is q^n. If the greatest carrier-to-noise ratio is selected in the required short interval,

$$p_n = \frac{q^n}{2} = \frac{1}{2(1+R)^n} \tag{1}$$

or approximately $1/2R^n$ for $R \gg 1$.

To find the diversity gain, let $p_1(R) = p_n(R_n)$. Then

$$1 + R = (1 + R_n)^n$$

and

$$G_n = \frac{R}{(1+R)^{1/n} - 1}, \tag{2}$$

or

$$G_n \approx R^{(n-1)/n} \tag{3}$$

for $R \gg n$.

Carrier Selection

In this method, the system selects the receiver containing the greatest carrier amplitude. The Rayleigh-distributed amplitude variation of any one carrier is specified by

$$y = \exp\left(-\frac{S^2}{S_0^2}\right)$$

where y is the probability that the rms carrier amplitude exceeds S, and S_0 is the long-term rms carrier amplitude. The relative probability that the greatest carrier amplitude is S is

$$-n(1-y)^{n-1}\frac{dy}{dS}.$$

The probability that demodulated output of this selected receiver is indeterminate[5] is y^R. If the greatest carrier is selected in the required short interval, then

$$\frac{dp_n}{dS} = -\frac{n}{2}(1-y)^{n-1}y^R\frac{dy}{dS}$$

where p_n is the n-diversity error, and

$$p_n = \frac{n}{2}\int_0^1 (1-y)^{n-1}y^R dy = \frac{n}{2}\frac{\Gamma(n)\Gamma(1+R)}{\Gamma(n+1+R)}$$

$$= \frac{n!}{2}\frac{\Gamma(1+R)}{\Gamma(n+1+R)} \tag{4}$$

in terms of the Gamma function.[6] Successive application of the formula

$$\Gamma(m+1) \equiv m\Gamma(m)$$

in the denominator reduces (4) to

$$p_n = \frac{n!}{2}\prod_{j=1}^{n}\frac{1}{j+R}. \tag{5}$$

Again, let $p_1(R) = p_n(R_n)$. Then, for $R \gg n$,

$$\frac{1}{R} \approx \frac{n!}{R_n{}^n}$$

and

$$G_n \approx \frac{R^{(n-1)/n}}{(n!)^{1/n}}. \tag{6}$$

Carrier-plus-Noise Selection

In this method, the system selects the receiver having the greatest voltage (or power) output. This criterion is the basis of diversity selection in many practical sys-

[6] L. A. Pipes, "Applied Mathematics for Engineers and Physicists," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 302–304; 1946.

tems. It has the advantage over the first two methods discussed in that there is no limitation on the speed of selection. It does not guarantee, however, that the selected receiver is the best one, either on a carrier-to-noise ratio or maximum-carrier basis, because the receiver in use may have been selected for large noise rather than large carrier.

Analysis of this system requires the joint amplitude probability function of carrier and carrier plus noise. Lacking this function, we can conclude only that the diversity error will be greater than that of carrier selection alone, (5), and that the diversity gain will be less than that of (6).

Dual-Filter Carrier-plus-Noise Selection

This system uses two narrow-band radio-frequency filters for each receiver, the two filters being centered at opposite limits of the frequency shift. The system selects the filter having the greatest output amplitude. The amplitude variation of any one filter output containing the carrier is specified by

$$y = \exp\left[-\frac{E^2}{(1+R)N_0^2}\right]$$

where y is the probability that the rms amplitude of the carrier and noise exceeds E, and N_0 is the long-term rms noise amplitude. The relative probability that the greatest carrier and noise amplitude is E is

$$-n(1-y)^{n-1}\frac{dy}{dE}.$$

The probability that the noise amplitude exceeds E in at least one filter output that contains no carrier is

$$1-(1-y^{1+R})^n.$$

Then,

$$\frac{dp_n}{dE} = -n(1-y)^{n-1}[1+(1-y^{1+R})^n]\frac{dy}{dE}$$

where p_n is the n-diversity error, and

$$p_n = n\int_0^1 (1-y)^{n-1}[1-(1-y^{1+R})^n]dy$$

$$= n\int_0^1 (1-y)^{n-1}\sum_{i=1}^{n}(-1)^{i+1}\frac{n!}{i!(n-i)!}y^{i(1+R)}dy.$$

In terms of the Gamma function,

$$p_n = (n!)^2\sum_{i=1}^{n}\frac{(-1)^{i+1}}{i!(n-i)!}\,\frac{\Gamma[1+i(1+R)]}{\Gamma[n+1+i(1+R)]}. \quad (7)$$

Application of the formula

$$\Gamma(m+1) \equiv m\Gamma(m)$$

reduces (7) to

$$p_n = (n!)^2\sum_{i=1}^{n}\frac{(-1)^{i+1}}{i!(n-i)!}\prod_{j=1}^{n}\frac{1}{j+i(1+R)}. \quad (8)$$

Let $p_1(R)=p_n(R_n)$. Then, for $R\gg n$,

$$\frac{1}{R} \approx \frac{(n!)^2}{R_n{}^n}\sum_{i=1}^{n}\frac{(-1)^{i+1}}{i!(n-i)!i^n}$$

and

$$G_n \approx \frac{R^{(n-1)/n}}{(n!)^{2/n}\left(\sum_{i=1}^{n}\frac{(-1)^{i+1}}{i!(n-i)!i^n}\right)^{1/n}}. \quad (9)$$

SIGNAL SELECTION

In this method, the system examines the n demodulated outputs. For binary transmission, some fraction, r, of the n receivers will indicate one signal, while the remainder, $n-r$, will indicate the other. The system must decide which signal is the correct one.

Let p be the probability that any one signal is incorrect. Then, the probability of any one combination of r incorrect signals is $p^r(1-p)^{n-r}$. The number of such combinations is

$${}_nC_r = \frac{n!}{r!(n-r)!}.$$

The total probability of exactly r incorrect signals is

$$q_r = {}_nC_r p^r(1-p)^{n-r}$$

and the probability of at least r incorrect signals is

$$p(r) = \sum_{i=r}^{n} q_i.$$

Let the system base its decision on a majority of the n signals; if the system is to render a decision for every possible combination, n must be odd. Then

$$r = \frac{n+1}{2}$$

and

$$p_n = \sum_{i=(n+1)/2}^{n}\frac{n!}{i!(n-i)!}p^i(1-p)^{n-i}$$

$$= n!p^n\sum_{j=0}^{(n-1)/2}\frac{1}{j!(n-j)!}\left(\frac{1-p}{p}\right)^j, \quad (n \text{ odd}). \quad (10)$$

For narrow-band FSK, the single-receiver error is

$$p = \frac{1}{2(1+R)}$$

and

$$p_n = \frac{n!}{2^n(1+R)^n}\sum_{j=0}^{(n-1)/2}\frac{(1+2R)^j}{j!(n-j)!}. \quad (11)$$

Let $p_1(R)=p_n(R_n)$. Then, for $R\gg 1$,

$$\frac{1}{2R} \approx \frac{n!}{\left(\frac{n+1}{2}\right)!\left(\frac{n-1}{2}\right)!(2R_n)^{(n+1)/2}}$$

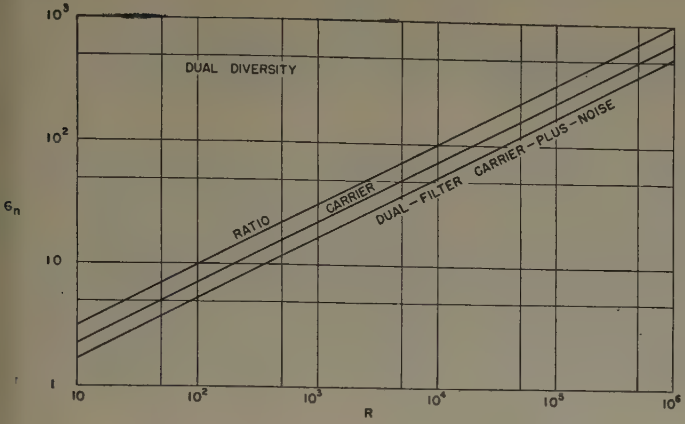

Fig. 2—Dual-diversity gain vs average carrier-to-noise ratio.

and

$$G_n \approx \left[\frac{\left(\frac{n+1}{2}\right)!\left(\frac{n-1}{2}\right)!}{n!} \right]^{2/(n+1)} (2R)^{(n-1)/(n+1)}. \quad (12)$$

Comparisons

Diversity gain equations (3), (6), and (9) are plotted in Fig. 2, assuming dual diversity. They are compared with (12) in Fig. 3, assuming triple diversity.

As an example, suppose that a particular narrow-band circuit requires a binary error no greater than 10^{-4} for satisfactory operation and that a transmitter carrier power of one kilowatt produces a carrier-to-noise ratio of 30 db at the receiver. Assume that the terminal equipment is ideal, that is, that

$$p = \frac{1}{2(1+R)}.$$

Then the necessary carrier-to-noise ratio for a single-receiver error of 10^{-4} is approximately

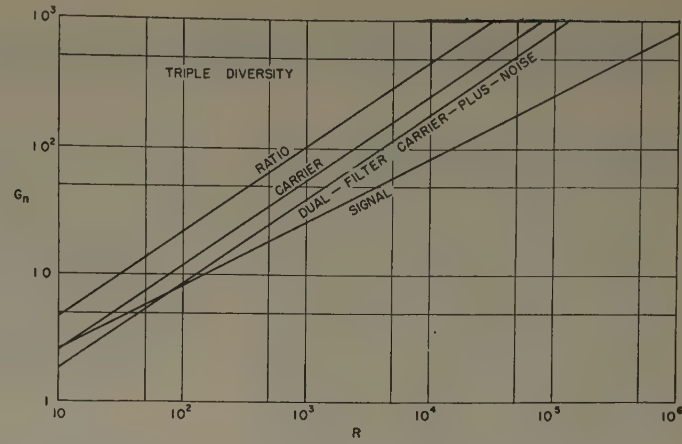

Fig. 3—Triple-diversity gain vs average carrier-to-noise ratio.

$$R = \frac{1}{2p} = 5 \times 10^3,$$

or 37 db. The following table shows the transmitter carrier power (in watts) required for an error of 10^{-4} using dual- and triple-diversity reception with each of the selection methods considered. Note that the figures for ratio- and carrier-selection diversity apply only for conditions of slow carrier fading.

TABLE I

Receiver	Selection Method				
	Ratio	Carrier	Carrier Plus Noise	Dual-Filter Carrier Plus Noise	Signal
Single	5,000	5,000	5,000	5,000	5,000
Dual-div.	71	100	>100	132	Not applicable
Triple-div.	15	31	> 31	43	87

Correspondence

Nonreciprocal Loss in Traveling-Wave Tubes Using Ferrite Attenuators*

Recent advances in waveguide devices making use of the phenomenon of ferromagnetic resonance, such as the microwave gyrator,[1] the microwave isolator,[2,3] and so forth, have led us to investigate how applications of the principle of nonreciprocity might benefit traveling-wave tubes and related devices.

Traveling-wave amplifiers[4,5] have a microwave circuit, commonly a helix, along which a signal can travel from the output to the input. Feedback due to reflections at the input and output can cause the tube to oscillate. Furthermore, the tube may oscillate as a backward-wave oscillator because of spatial harmonics of the circuit wave which travel in the direction of electron flow when the wave travels in the opposite direction.[6,7,8]

Such oscillations can be avoided with some reduction in gain by introducing loss along the circuit, either a moderate uniform loss or a very high loss over a short length somewhere between the input and the output. Concentrated (or "lumped") loss does not necessarily eliminate backward-wave oscillations; it can lead to severe departures from uniformity of gain versus frequency, and it may reduce tube efficiency.[9] Sufficient uniform loss reduces both gain and efficiency considerably.

For these reasons, and some others that have not been mentioned, one would wish to have a circuit which has little loss in the direction of electron flow and large, uniformly distributed loss in the opposite (backward) direction.

The nonreciprocal behavior of ferrites associated with the phenomenon of ferromagnetic resonance absorption makes it possible to fulfill this wish. The rf magnetic-field distribution surrounding a helix is such as to lead to very simple constructions of isolators, i.e., devices in which there is little loss in the forward direction and substantial loss in the backward direction.

Fig. 1 shows a sketch of the distribution of rf magnetic lines of force inside and outside a helix. If we project the magnetic vector at P, say, onto a plane through the axis

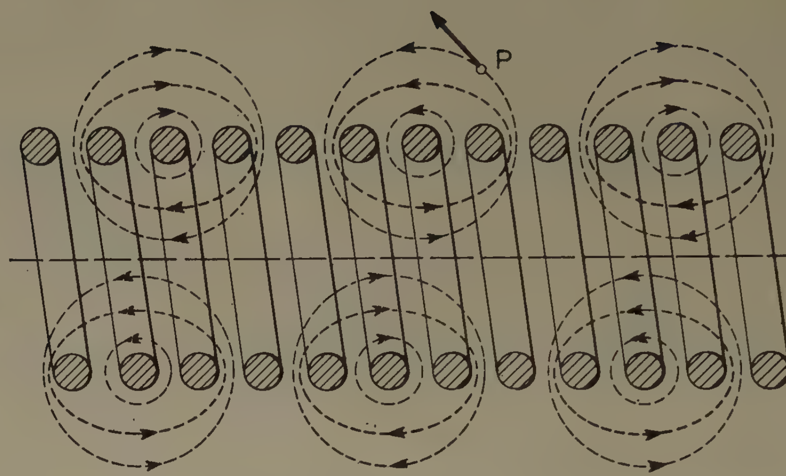

Fig. 1—Sketch of magnetic field distribution around a helix.

it will be seen that this vector rotates counterclockwise for a wave traveling from left to right, or clockwise for a wave traveling from right to left. It can be shown, moreover, that over most of the space surrounding the helix the projection of the amplitude stays very nearly constant. In this region therefore the magnetic field can be regarded as nearly circularly polarized. As has been shown by Miller, et al. and by Kales, et al. this is a situation favorable for nonreciprocal interaction. This is indeed what happens: The combination of helix and circumferentially magnetized ferrite acts just like a waveguide isolator—giving low loss in the forward direction and high loss in the backward direction.

Experiments in the 4,000-mc range showed that the desired kind of interaction can be readily obtained with ferrites of the nickel-zinc type. Better than 10 to 1 backward-to-forward loss ratios are easily obtained, the actual magnitude of the loss depending on the distance between the helix and the ferrite. The ferrite cylinders used in these experiments were magnetized by passing a few amperes through a few turns of wires linking the ferrite.

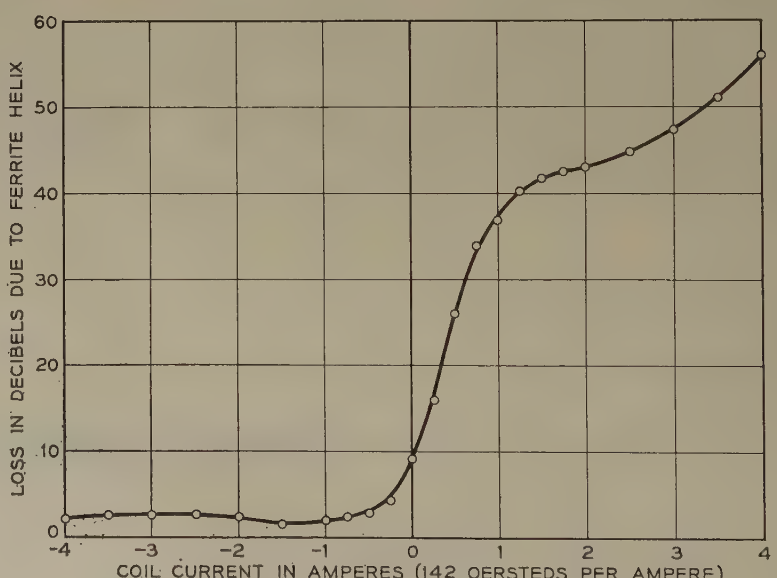

Fig. 2—Distributed loss of a helix surrounded by a helical ferrite (#2010) (3 inches long, 10 turns per inch, 0.500-inch outer diameter, 0.210-inch inner diameter) as a function of applied magnetic field. The measurement was done at 4,200 mc.

interaction between an rf field and a ferkite, magnetized in a direction perpendicular to the plane of the rf magnetic field; a cylinder of ferrite surrounding the helix with a circumferential component of magnetization would be expected to produce nonreciprocal

* Received by the IRE, May 7, 1954.

[1] C. L. Hogan, "Ferromagnetic Farraday effect at microwave frequencies and its applications—the microwave gyrator," *Bell Sys. Tech. Jour.*, vol. 31, p. 1; 1952.

[2] M. L. Kales, H. N. Chait, and N. G. Sakiotis, "Non-reciprocal microwave components," *Jour. Appl. Phys.*, vol. 24, p. 816; 1953.

[3] S. E. Miller, A. G. Fox, and M. T. Weiss, *Bell Sys. Tech. Jour.*; to be published in the near future.

[4] J. R. Pierce, "Traveling Wave Tubes," Van Nostrand and Co., New York, N. Y.; 1950.

[5] R. Kompfner, "Travelling wave tubes," *Rep. Prog. Phys.*, vol. 15, p. 275; 1950.

[6] P. K. Tien, "Traveling-wave tube helix impedance," Proc. I.R.E., vol. 41, p. 1617; November, 1953.

[7] S. Sensiper, "Electromagnetic Wave Propagation on Helical Conductors," Mass. Inst. Tech. thesis; 1951.

[8] R. Kompfner and N. T. Williams, "Backward-wave tubes," Proc. I.R.E., vol. 41, p. 1602; November 1953.

[9] C. C. Cutler and D. J. Brangaccio, "Factors affecting travelling wave tube power capacity," *Trans. I.R.E. PGED-3*, p. 9; June, 1953.

Correspondence

The electron beam in traveling-wave tubes is usually focused by an axial field of several hundred oersted. A long cylinder of ferrite surrounding the helix would not only severely reduce the field inside but would cause mainly reciprocal interaction with waves on the helix when immersed in such an axial field. Therefore it is necessary to divide the ferrite into a succession of relatively narrow rings spaced apart with a spacing approximately equal to the thickness of the rings. It is then found that the axial field is substantially the same as would exist in the absence of the ferrite rings, while the effective longitudinal component of magnetization inside the ferrite rings can be made negligible compared with the circumferential component produced by means of the current in the windings interlinking the rings.

If the ferrite itself is formed into a helix, surrounding the metallic rf power-carrying helix, the focusing field will itself produce a circumferential component of magnetization in the ferrite. This eliminates the need for a winding linking the ferrite and permits a simple construction of isolator particularly adapted to helix-type traveling-wave tubes in which the beam is focused with an axial field.

Figs. 2, on the previous page, and 3 show the additional forward and backward loss caused by such a ferrite helix surrounding a metallic helix of 0.135-inch diameter and 20 turns per inch, as a function of magnetic field, at a fixed frequency of 4,200 megacycles and as a function of frequency at a fixed field of 425 oersted, respectively.

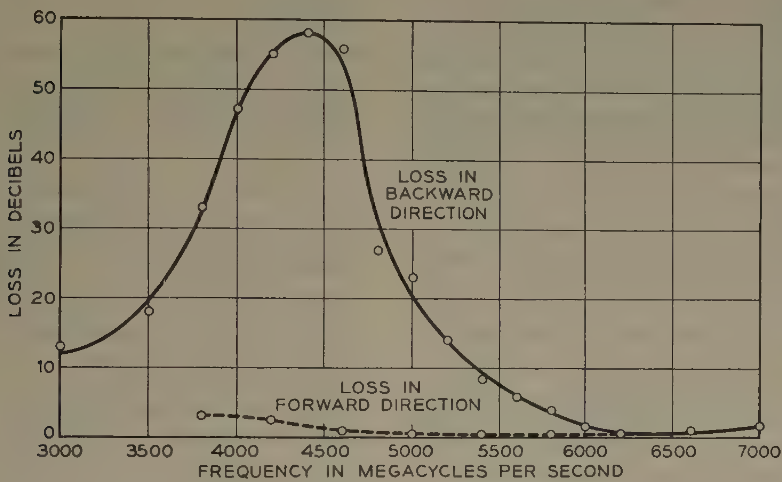

Fig. 3—Distributed loss of a helix surrounded by a helical ferrite (#2010) (similar dimensions as in Fig. 2) at a constant applied field of 425 oersted, as a function of frequency.

Preliminary experiments with traveling-wave tubes show that this kind of non-reciprocal loss does indeed allow one to obtain gain substantially exceeding the forward loss.

It is a pleasure to acknowledge our indebtedness to a number of people for instruction, advice and practical help in this work, in particular to C. L. Hogan, J. H. Rowen, and L. G. Van Uitert.

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R. Kompfner, and
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Murray Hill, N. J.

False Alarm Time in Pulse Radar*

In pulse-radar-system studies, the concept of a false alarm time is often found useful since it measures the rate at which false alarms, or noise pulses that are interpreted as target echoes, occur. Unfortunately, there exist at least two different definitions of this time interval in the literature.

Perhaps the most common definition of false alarm time is that it is the average time interval between false target indications.[1] To determine the probability that one false alarm will occur within the false alarm time as defined above, we may apply the binomial distribution since the probability of a false alarm occurring at any given instant is constant. However, since this probability is so small, and the number of possible successes is so great, it will be more convenient to use Poisson's distribution, which is the limiting case of the binomial distribution.

This distribution is

$$P(r) = \frac{m^r e^{-m}}{r!},$$

with $P(r)$ as the probability of r successes, and m the average or expected number of successes. In accordance with the above definition of false alarm time, $m=1$. The probability of one false alarm within this time interval is $e^{-1} \doteq 0.368$. The probability of other number of false alarms may be similarly calculated.

Another definition of false alarm time appears in the literature.[2] According to this definition, a false alarm time is defined as the time in which the probability is $\frac{1}{2}$ that a false alarm will not occur. The Poisson distribution may again be applied with $P(0)=1/2$. Thus

$$1/2 = e^{-m}.$$

The average number of false alarms occurring within this time interval is $m=0.693$. The probability of one false alarm in this interval is

$$P(1) = 0.693e^{-0.693} \doteq 0.346.$$

Thus the false alarm time as defined by Kaplan and McFall is about 45 per cent longer than Marcum's alarm time although the probabilities of one false alarm within their respective alarm times differ by only about 6 per cent.

The differences between these two definitions can perhaps best be illustrated by Table I, shown below:

TABLE I

r	$P(r)$ K-McFall Mean=1	$P(r)$ Marcum Mean=0.693
0	0.368	0.500
1	0.368	0.346
2	0.184	0.120
3	0.061	0.028
4	0.015	0.005

Richard Hollis
Bendix Radio, Communications Div. of
Bendix Aviation Corp.
Baltimore, Md.

* Received by the IRE, February 8, 1954.

[1] S. M. Kaplan and R. W. McFall, "The statistical properties of noise applied to radar performance," Proc. I.R.E., vol. 39, pp. 56–60; January, 1951.

[2] J. I. Marcum, Douglas Aircraft Co., Inc., Report No. RA-15061.

Correspondence

Dr. Sterky on Filters*

On page 777 of the June, 1953 issue of the PROCEEDINGS Strasberg has an article entitled, "The Effects of Terminations and Dissipation on the Insertion Loss of Some Simple Ladder Filters." Although my present duties give me little time to study recent articles in different journals on filter problems, I have as a former professor of telegraphy and telephony at the Royal Technical Institute of Stockholm still a special liking for subjects close to that of my doctor's thesis. I believe that in my thesis, "Methods of Computing and Improving the Complex Effective Attenuation, Load Impedances and Reflexion Coefficients of Electric Wave Filters," which was presented on May 18th, 1933, I introduced a concept, the so-called α-matching, which covers the case on a principal basis when the dissipations in the series arms (or the shunt arms) are represented by resistances as part of the mid-series (or mid-shunt) terminations of ladder filters. This thesis also appeared in the *Ericsson Technics*, number 4, 1933.

It might also be of interest to point out that the problem of computing the effective attenuations or insertion losses of ladder filters has been extensively treated in many articles of Swedish origin, some of them printed also in French, English, and German. The book "Fyrpolteorier och frekvenstransformationer" (The Theory of Quadripoles and Frequency Transformations) by my successor as professor of telegraphy and telephony of the Swedish Royal Technical Institute of Stockholm, Torbern Laurent, contains in chapter 9 a complete list of references of articles which you—if I may say so—seem to overlook in the United States.

I hope that my observations will be accepted not as criticism but only as a contribution to the universality of science and engineering.

HÅKAN STERKY
Director General
Swedish Telecommunication
and Administration
Stockholm, Sweden

* Received by the IRE, July 9, 1953.

Reply to Dr. Sterky on Filters*

It was a pleasure to receive Dr. Sterky's letter and a copy of his thesis.

My article[1] is based on some work with filters done in 1945. Upon reviewing an old notebook of mine, some of the material seemed worth publishing at this time, principally for two reasons: Some of the information did not appear to have been published before; it seemed to me that many people using filters were not aware that nonstandard terminations are often preferable to termination with nominal design resistance.

Dr. Sterky's dissertation treats the effects of terminations in a complete and elegant manner, and I would most certainly have referenced it had I been aware of it. It is unfortunate that I was not previously aware of his work for it would have saved me much time thinking about the subject in general. Specifically, I was unaware of the beautifully compact expression for the insertion loss given as his equation (16), and unaware of the publication of curves of either "effective attenuation" or insertion loss for terminations which differ from the nominal design resistance, which he calls "α-matching," other than those in the references cited in my article.[2]

However, I believe that my article does present several items previously unpublished, viz:

(1) The use of a frequency parameter F and dissipation parameter D defined in such a way that the frequency parameter is independent of dissipation and the dissipation parameter is independent of signal frequency. The parameters U and V, originated by Zobel, which Sterky uses in his treatment of the effects of dissipation, do not have this independence and are therefore less convenient. Others have used parameters which coincide with mine for low- and high-pass filters, and with my approximate form for narrow-band filters, but the exact form applying to band filters of any bandwidth I have not seen published before.

(2) The publication of curves of insertion loss indicating the *combined* effects of dissipation and of termination. Sterky's article considers these effects separately.

(3) The inclusion of the dissipation in the termination arms as part of the terminating resistances is different from "α-matching" since it permits the design of filters with greater dissipation in terminating arms than in intermediate sections.

Sterky's comment that we in the United States tend to overlook publications from other countries in this field is well taken. This perhaps is partly caused by our difficulty with other languages, a difficulty, which many of his countrymen do not have as his letter in English so well exemplifies.

I am grateful to Dr. Sterky for bringing his thesis and the other references to my attention.

M. STRASBERG
Navy Department
David Taylor Model Basin
Washington, D. C.

* Received by the IRE, October 19, 1953; revised December 16, 1953.

[1] M. Strasberg, "The effects of terminations and dissipation on the insertion loss of some simple ladder filters," PROC. I.R.E., vol. 41, pp. 777–780; June, 1953.

[2] The use of nonnominal resistance terminations is also treated in a book by J. H. Mole, "Filter Design Data for Communication Engineers," E. & F. N. Spon Ltd., London, England; 1952. Mole calls this "ρ matching." This book came to my attention subsequent to the preparation of my article.

Standards on Sound Recording and Reproducing*

The *IRE Standards on Sound Recording and Reproducing: Methods of Determining Flutter Content*[1] was approved by the American Standards Association on March 16, 1954, as the *American Standard, Method for Determining Flutter Contents of Sound Recorders and Reproducers, Z57.1-1954*. This announcement marked the conclusion of some eight years of efforts to achieve a flutter standard.

The IRE technical committees and their members who worked on this Standard were duly noted on page 537 of the March issue. However, those who use this Standard may wish to know also of the important contributions made by SMPTE and ASA committees.

This Standard was first developed by the Sound Committee of the Society of Motion Picture and Television Engineers under the chairmanship of John G. Frayne in 1946 and was published for comment as a proposed standard in the August, 1947 issue of the SMPTE Journal. As a result of this publication, rather extensive comments were received from all fields of sound recording.

In October, 1947 the ASA set up a Sectional Committee on Sound Recording, Z57, sponsored jointly by IRE and SMPTE and this proposed Standard was referred to it for processing as an American Standard. In view of the comments the SMPTE had received, a subcommittee of Z57, under the chairmanship of Dr. E. W. Kellogg, was set up to consider these suggested modifications. During the next five years, several drafts were drawn up by this subcommittee and comments were received from appropriate committees of the IRE, SMPTE, RETMA, and ASA. The final draft was approved by the ASA Sectional Committee Z57 in February 1952. It was subsequently approved by both sponsors of Z57, i.e., the SMPTE and the IRE and, as noted above, has now become an American Standard.

During the period 1948–1952 the SMPTE Sound Committee under the chairmanship of Lloyd Goldsmith played an important role in the development of this Standard. Credit should also be given to William H. Deacy, Staff Engineer of the SMPTE during this period, for his considerable labor in shepherding this project through its manifold procedures and channels.

AXEL G. JENSEN
Chairman,
IRE Standards Committee

* Received by the IRE, April 9, 1954.

[1] "Standards on Sound Recording and Reproducing: Methods for Determining Flutter Content, 1953," PROC. I.R.E., vol. 42, pp. 537–541; March, 1954.

Correspondence

Comments on "Figure of Merit for Communication Devices"*

C. E. Shannon has shown[1] that for a given signal power P and constant noise spectral density N_0 the capacity of a channel is maximized if the bandwidth is allowed to approach infinity. This maximum value is given by

$$C_{max}/P = \frac{1}{N_0}\log_2 e \quad \text{bits/second per watt of signal power.}$$

Based on this result, Felker[2] has made the very interesting observation that since in any physical system the minimum value of N_0 is equal to the thermal noise power density kT, a definite lower limit is set to the "cost" of information; from the above expression it is seen to be

$$E_{\min} = N_{0_{\min}} \ln 2 = kT \ln 2 \quad \text{joules per bit,}$$

or slightly less than 3×10^{-24} joules for T near room temperature. Felker concludes that a repeater or amplifier that is to pass intelligence at the rate of q bits/second must therefore have a battery drain of at least $3\times10^{-24}\,q$ watts. He then gives an example of a hypothetical transistor amplifier drawing one microwatt of power for which he computes a "figure of merit" (ratio of minimum to actual battery drain) of about 10^{-9}.

The purpose of this note is to point out (1) a confusion of concepts in Felker's article, (2) a proper definition of a figure of merit, and (3) the fact that for a practical wide-band communication device such as a pulse radar the value of this figure does not fall below unity by more than about two (rather than nine) orders of magnitude.

As to the first point, it is merely necessary to recall that channel capacity is a function of *signal* power. Any power expenditure in the amplification process, such as for plate and filament supply, is entirely irrelevant from an information theoretical point of view (except inasmuch as part of it appears as tube noise and thereby actually reduces the possible intelligence rate).

A proper figure of merit for a communication device can still be defined, though not in terms of battery drain, but rather as the ratio of E, the *signal input energy* required per unit of intelligence at the output, to $E_{\min}$, the above found minimum value of $kT \ln 2$ joules per bit. An entirely equivalent definition could also be made in terms of signal power instead of signal energy and intelligence rate in lieu of intelligence.

As a simple example, consider the problem of target range determination by means of an A-scope radar. The information gain, in bits, will be equal to the difference between H_1, the information function ("entropy") corresponding to p_1, the a priori probability distribution of the range, and H_2 the corresponding quantity for p_2, the a posteriori range distribution. Assuming p_1 to be a uniform distribution,

$$p_1 = \frac{1}{R_0} \qquad 0 \leq R \leq R_0,$$

where R_0 is the maximum range, gives

$$H_1 = \int p_1(R)\log_2 p_1(R)dR = \log_2 R_0.$$

Supposing further that the signal-to-noise ratio and number of sweeps are sufficiently high so that range can be read off with an error having a distribution[3] with $\pm 2\sigma$ points at the pulse edges, then

$$p_2(R) = \frac{1}{\sqrt{2\pi}\,\sigma}e^{-R^2/2\sigma^2} \qquad \sigma = \frac{c\tau}{8},$$

where τ is the pulse duration; hence

$$H_2 = \int p_2(R)\log_2 p_2(R)dR$$

$$= \log_2\left(\sqrt{2\pi e}\,\frac{c\tau}{8}\right) \simeq \log_2\frac{c\tau}{2}.$$

The information gain is therefore

$$Q = H_1 - H_2 \simeq \log_2\frac{2R_0}{c\tau} \qquad \text{bits.}$$

Typical values of Q may range between 5 and 10. The signal energy required for the above range accuracy with conventional A-scope presentation may be estimated from photographs of A-scope displays:[4] between 25 and 50 sweeps at a ratio of signal to average noise power of 3 appear sufficient. The corresponding energy is approximately

$$50S\tau = 50 \times (3 \times \text{noise figure} \times kTW)\tau$$

$$= 1{,}500kT$$

where S is the signal-pulse power and W, the IF bandwidth, is taken equal to the reciprocal pulse length; a noise figure of 10 db has been allowed for. It is seen then that E for this device is around 200 kT joules per bit as compared to the minimum value of $(\ln 2)kT$, corresponding to a figure of merit of the order of 10^{-2}. This calculation is admittedly rather crude (it would require near targets to be small and vice versa, since no account of has been taken of the dependence of signal strength on range), but the order of magnitude is believed to be fairly typical. A similar result is obtained, for example, for a receiving set receiving carrier-modulated binary PCM. The signal-to-noise ratio required in this case for negligible equivocation is about 12 db (16:1).[5] The energy per pulse is therefore

$$16(N_0W)\tau = 16 \times \text{noise figure} \times kT$$

assuming $W\tau = 1$. Since for zero equivocation each pulse conveys one bit of information, it is seen that $E \cong 16\times\text{n.f.}\times kT$, yielding a figure of merit $(E/E_{\min})$ of about the same magnitude as found in the first example.

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Hughes Aircraft Co.
Culver City, Calif.

* Received by the IRE, February 23, 1954.

[1] C. E. Shannon, "Communication in the presence of noise," PROC. I.R.E., vol. 37, pp. 10–21; January, 1949.

[2] J. H. Felker, "A link between information and energy," PROC. I.R.E., vol. 40, p. 728; June, 1952.

[3] The error distribution is taken to be Gaussian, with variance σ^2.

[4] Using Fig. 2.7, "Radar System Engineering," MIT Rad. Lab. Series, vol. 1, p. 42; 1947.

Rebuttal[6]

Upon reading the above communication, I was reminded of an acquaintance who bought a ticket to Albany, got on a bus, discovered that it was going to California, decided that the driver was confused and undertook to teach him the route to Albany. Just as there are many legitimate destinations for a bus, there are many figures of merit. The figure of merit proposed by F. P. Adler will no doubt be a useful measure of coding efficiency and noise figure.

When I wrote my original letter I was, to use the bus analogy, not interested in going to Albany. The ground had been fairly well covered before.[7] My interest was first of all to show what still appears to me to be a very neat way of linking information and energy.[8] Having obtained an exchange rate between energy and information, my next step was to suggest that we relate the energy equivalence of the information a circuit processes to the energy required to operate the circuit. If our bookkeeping is to be complete, we must include "battery" energy as well as signal energy in the cost of operating the circuit. In fact, the "battery" energy generally exceeds the signal energy to such an extent that the signal energy can be neglected in computing a figure of merit. I recognize that it may be desirable to subtract the energy output of the circuit from the "battery" energy before computing the figure of merit.

The figure of merit that I proposed is not only of philosophical interest but will become of increasing practical significance as more efficient active elements become available. It may not be useful in all fields and my definition may require amendment or correction. I will be glad to hear of efforts to sharpen it. My guess is that such contributions will leave the destination of the hypothetical bus in about the same direction as it originally started.

J. H. FELKER
Bell Telephone Laboratories, Inc.
Whippany, N. J.

[5] B. M. Oliver, J. R. Pierce, and C. E. Shannon, "The philosophy of PCM," PROC. I.R.E., vol. 36, pp. 1324–1331; November, 1948.

[6] Received by the IRE, March 15, 1954.

[7] For example, P. M. Woodward and I. L. Davis, "A theory of radar information," *Phil. Mag.*, vol. 41, p. 1001; October, 1950. In this study it was shown that "information is initially obtained at a roughly uniform rate which is not far removed from the absolute limit for an ideal communications system."

[8] For a different path to the same result see L. Brillouin, "The negentropy," *Jour. Appl. Phys.*, vol. 24, p. 1152; September, 1953.

Correspondence

Silicon Transistor*

The larger energy gap of silicon as compared to germanium has led to widespread interest in its possible use in transistors. It is known that higher operating temperatures are possible; however, the temperature dependence of the operating parameters for silicon transistors has not been published. The *n-p-n* silicon transistors whose characteristics are reported below were made from silicon single crystals with base layers approximately 0.0005 inch in thickness. Base contacts were made by alloying aluminum wire into the base region. Minority-carrier lifetimes in the collector region were about 20 microseconds because of adding donor impurities to control collector resistivity.

Typical characteristics of these silicon transistors compared with a representative germanium transistor are given in Table I, below right.

Examples of the temperature dependence of the current amplification factor for grounded base connection are shown in Fig. 1. Fig. 2 shows the variation of I_{c0} with temperature for some representative units. It is also of interest how the transistor parameters r_b, r_c, and r_e vary with ambient temperature. Average variations for a number of units are given in Table II, below right.

The power-handling capabilities of silicon transistors are considerably higher than germanium units mounted and enclosed in the same fashion. At an ambient temperature of 25 degrees C, these silicon junction transistors are capable of dissipating 250 milliwatts. At ambient temperatures of 100 degrees C, this is reduced to about one half this value. By the simple expedient of attaching the transistor case to the chassis, typical silicon units will handle powers in the 1.0- to 1.5-watt range at ambient temperatures of 25 degrees C. If the ambient temperature is raised to 100 degrees C, this value is reduced also by about one half.

In Fig. 3 are shown plots of the current-voltage characteristics for the emitter *p-n* junction of a representative germanium transistor and for a typical silicon transistor. These curves show why it is necessary to bias the emitter of a silicon *n-p-n* junction transistor farther in the forward direction than is the practice with germanium units. This bias condition implies that circuits employing germanium transistors will have to be modified somewhat to adapt them for silicon transistor use.

The above data show that the silicon *n-p-n* grown junction transistor is indeed a device capable of extending the temperature range of transistors to something comparable with other circuit components. It shows exceptional promise for high-frequency applications, for uses where high-power dissipation is needed, and for conditions necessitating high ambient temperatures.

W. A. Adcock
M. E. Jones
J. W. Thornhill
E. D. Jackson
Texas Instruments Incorporated
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* Received by the IRE, June 19, 1954

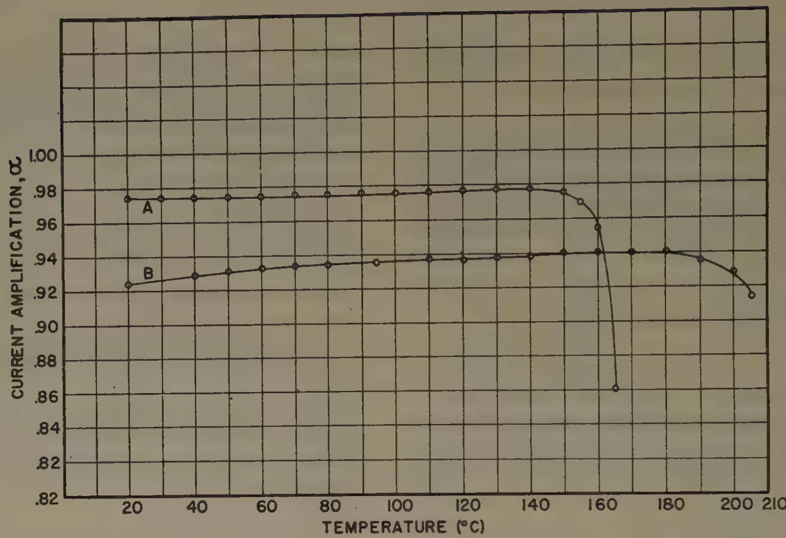

Fig. 1—Variation of current-amplification factor with temperature for grown silicon junction transistors.

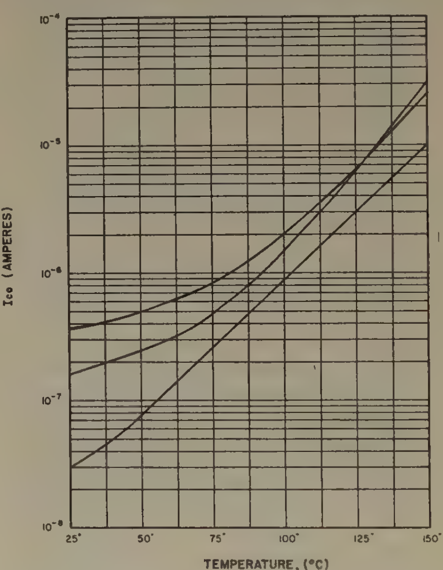

Fig. 2—Variation of I_{c0} with temperature for grown silicon junction transistors. (Collector voltage 5 volts.)

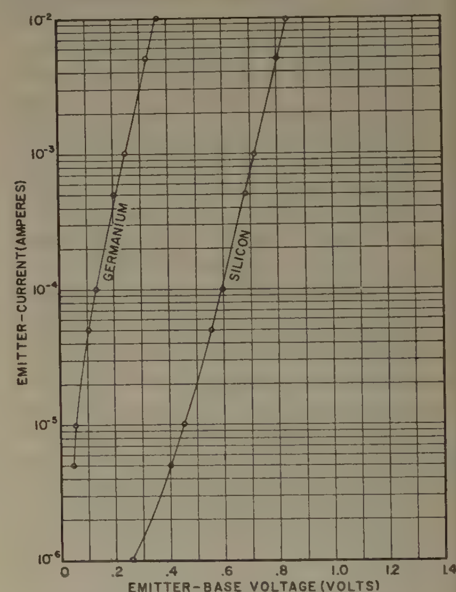

Fig. 3—Current-voltage characteristics for emitter *p-n* junction.

TABLE I

ELECTRICAL DATA ON SILICON TRANSISTORS COMPARED WITH A REPRESENTATIVE GERMANIUM TRANSISTOR

Parameter	Silicon Transistor	Germanium Transistor
I_{c0} at 25°C (V_c=5.0 v.)	≦0.4 μamp	4.0 μamp
I_{c0} at 150°C (V_c=5.0 v.)	20–50 μamp	very high
r_e	25–75 ohms	22 ohms
r_b	100–300 ohms	170 ohms
r_c	0.6–1.5 megohms	1.8 megohms
Current Amplification Factor* (α)	0.90–0.98	0.96
Frequency Cut-off	1.2–>7.5 mc	1.1 mc
Noise Figure	16–22 db	23 db

* Grounded Base Connection.

TABLE II

VARIATION OF TRANSISTOR PARAMETERS WITH TEMPERATURE

Parameter	25°C	150°C	Change
r_c	1.34 megohms	0.27 megohms	−1.07 megohms
r_b	231 ohms	928 ohms	+697 ohms
r_e	54 ohms	34 ohms	−20 ohms

Correspondence

Audio Amplifiers*

About two years ago an article by Peterson and Sinclair[1] demonstrated a method for eliminating switching transients in Class AB and B amplifiers. The idea was extremely valuable and a great contribution to the field. However, the circuit contains an unbalanced output stage, making it difficult to drive. This correspondence describes a balanced push-pull circuit without the switching transient.

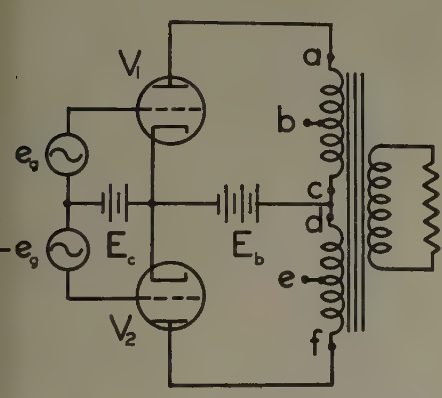

Fig. 1—Conventional push-pull circuit.

The switching transient is caused by the leakage inductance between the two sections of the primary winding of the output transformer.[2] It cannot be avoided in Class AB and B amplifiers where the plate current is cut off during a part of the cycle. To eliminate this transient, it is necessary either to have a common load impedance for both output tubes or to reduce the leakage inductance to a negligible value. The circuit described here uses the first approach.

In Fig. 1 a conventional push-pull circuit is shown. If b and e are the center-taps of the two sections ac and df of the primary winding, then the transformer may be connected as in Fig. 2 with the same load impedance to the output tubes. Since, in Fig. 2, a and d and c and f are at the same ac potentials respectively, the condensers C_1 and C_2 may be inserted without upsetting the push-pull operation. If the capacitances of C_1 and C_2 are reasonably large, V_1 and V_2 may be considered to have a common load, which is the condition for the elimination of switching transients, and yet the circuit is balanced.

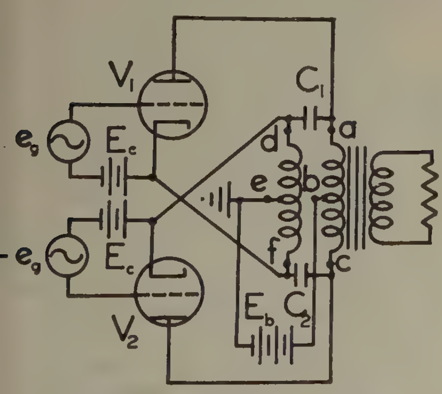

Fig. 2—New push-pull circuit.

The same result may be achieved by using two conventional output transformers, one in the plate circuit and the other in the cathode circuit. It is necessary to terminate the secondaries, properly phased, in a common load. This is not recommended, however, because of the added cost and the reduction in primary inductance by a factor of two.

As for the size of C_1 and C_2, it is sufficient if the reactance is negligible at the high frequencies where the leakage inductance between ac and df presents an appreciable reactance.

Since the leakage inductance between the sections of the primary is not critical in this circuit, the cost of the transformer will be reduced. Of course, the leakage inductance between the primary and the secondary must be small and the primary inductance must be large when a high-quality amplifier is desired.

So far, the problem of driving has not been considered. Usually, the driving signals are applied between the grid and the ground instead of between the grid and the cathode as in Fig. 2. This means that the operation is similar to that of a cathode follower. However, it may be noted that the required driving voltage is about one-half of that for the cathode follower for the same power output.

Taps b and e may be shifted keeping the same load condition for V_1 and V_2. In other words, b may be shifted toward a provided that e is shifted toward d by the same amount. Of course, this shift destroys the balance, but not the load condition. As the extreme, b and e are brought to a and d. The result is one of the circuits of Peterson and Sinclair.

Pentodes or beam-power tetrodes, instead of triodes, may well be used in the circuit with the screen of V_1 connected to c and the screen of V_2 connected to a. If the screens are connected to b, then the operation is somewhat like that of the ultra-linear circuit.[3]

The same idea may be applied to a push-pull circuit where the load is split between the plate and the cathode but the number of turns of these windings are not equal.[4] In Fig. 3 a circuit is shown where the number of turns of the winding df is smaller than that of the winding ac. In this case, C_1 and C_2 are connected from d and f to a' and c' respectively, where the number of turns of de, ef, $a'b$, and bc' are equal. With these condensers, the leakage inductance between the primary windings which contributes to the generation of switching transients is that between aa' and cc'. In other words, if the transformer is wound so as to minimize the leakage between aa' and cc', then ths switching transients can be eliminated. It ie obvious that a transformer may be wound more cheaply this way than by trying to reduce the leakage between the whole sections of the primary.

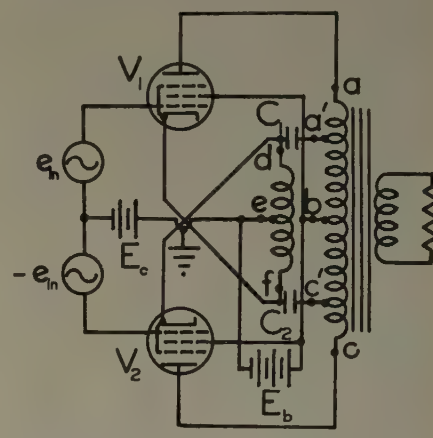

Fig. 3—Modified form of the new push-pull circuit.

The author admits that the circuits described here are not readily adaptable to operation without output transformer, which is one of the features of the circuit by Peterson and Sinclair. However, he will be glad to have a transformer in exchange for the absence of switching transients.

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Cornell University
Ithaca, New York

A Note on Information Theory*

Several proofs of the nonnegativeness of Shannon's information function

$$I = \sum_{i,j} p(i,j) \log \frac{p(i,j)}{p(i)p(j)}$$

have been given in the literature.[1,2] The following simple proof seems not to be generally known.

Since the curve $y = \log x$ is convex, it lies below its tangent at $x = 1$, and hence $\log x \leqq x - 1$ for all $x > 0$, with equality only for $x = 1$. Hence

$$\begin{aligned} -I &\leq \sum_{i,j} p(i,j) \left[\frac{p(i)p(j)}{p(i,j)} - 1 \right] \\ &= \sum_{i,j} p(i)p(j) - \sum_{i,j} p(i,j) \\ &= 0, \end{aligned}$$

equality holding only if $p(i,j) = p(i)p(j)$ for all i, j. It will be observed that the same result is obtained when $p(i)p(j)$ is replaced by any nonnegative function $q(i,j)$ such that $\sum_{i,j} q(i,j) = 1$.

H. Robbins
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Columbia University
New York 27, N. Y.

* Received by the IRE, February 11, 1954.
[1] A. Peterson and D. B. Sinclair, "A single-ended push-pull audio amplifier," Proc. I.R.E., vol. 40, pp. 7–11; January, 1952.
[2] A. Pen-Tung Sah, "Quasi transients in Class B audio-frequency push-pull amplifiers," Proc. I.R.E., vol. 24, pp. 1522–1541; November, 1936.
[3] D. Hafler and H. I. Keroes, "An ultra-linear amplifier," *Audio Eng.*, vol. 35, no. 11, pp. 15–17; November, 1951.
[4] D. T. N. Williamson and P. J. Walker, "Amplifiers and superatives," *Wireless World*, vol. 58, pp. 357–361; September, 1952.

* Received by the IRE, January 11, 1954.
[1] S. Goldman, "Information Theory," Prentice-Hall, Inc., New York, N. Y.; 1953.
[2] A. F. Laemmel, "General Theory of Communication," Polytechnic Institute of Brooklyn, New York, N. Y.; 1949.

IRE News and Radio Notes

Nominations—1955

At its May 5, 1954 meeting, the IRE Board of Directors received the recommendations of the Nominations Committee and the reports of the Regional Committees for officers and directors for 1955. They are:

President, 1955: John D. Ryder
Vice President, 1955: Franz Tank
Director-at-Large, 1955–1957 (*two to be elected*): John F. Byrne, Edward W. Herold, Ernst Weber, Jerome B. Wiesner
Regional Directors, 1955–1956 (*one to be elected in each Region*):
Region 2.—John N. Dyer
Region 4.—Semi J. Begun, E. M. Boone, Howard R. Hegbar
Region 6.—Durward J. Tucker
Region 8.—John T. Henderson

According to Article VI, Section 1, of the IRE Constitution, nominations by petition for any of the above offices may be made by letter to the Board of Directors, giving name of proposed candidate and office for which it is desired he be nominated. For acceptance a letter of petition must reach the executive office before noon on August 13, 1954, and shall be signed by at least 100 voting members qualified to vote for the office of the candidate nominated.

WESCON Plans Nearing Completion

The eyes of the nation's electronic industry will be on the "City of the Angels" August 25–27, when the 1954 WESCON gets underway in Los Angeles' Pan-Pacific Auditorium and Ambassador Hotel.

WESCON will be sponsored jointly by WCEMA (West Coast Electronic Manufacturers Association) and the Los Angeles and San Francisco Sections of the I.R.E. WESCON is recognized as the electronic industry's second largest annual event; only the I.R.E. National Convention exceeds it in size. According to W. D. Hershberger, Chairman of the WESCON Board of Directors, this year's Show and Convention is expected to surpass all previous Western records, in both attendance and exhibitor participation.

Paralleling the fabulous growth of the West in recent years, this year's event has exceeded all predictions, and is literally "bursting at the seams." Already more than 500 exhibit booths have been reserved—compared to a total of 370 occupied in last year's Show. It has been necessary to add an 11,000-square foot annex to the Pan-Pacific Auditorium in order to accommodate exhibitors desiring space.

Nearly 20,000 leaders of the nation's electronic industry and IRE members are expected to attend the three-day meeting. Work has been underway since the first of the year, under the leadership of C. F. Wolcott to make the Convention profitable and enjoyable for everyone.

Twenty-eight technical sessions are on the program, featuring carefully selected high-level papers presented by some of the nation's leading engineers and scientists. Tentative plans call for sessions on audio, antennas and propagation, circuit theory, vehicular communications, broadcast and TV, telemetering, airborne electronics, information theory, management, electron devices, computers, microwave theory, and component parts.

Sessions and panels are arranged in a general schedule of ten sessions per day with additional sessions of special interest in the evenings. More than 100 technical papers will be presented in all. Several outstanding special events (see Proc. I.R.E., June, 1954, page 1028) will also be held, with the All-Industry Luncheon the high spot of the social program. The featured speaker at the Luncheon will be William R. Hewlett, National President of the I.R.E. WCEMA Scholarship Awards will be presented at this time to outstanding students of accredited Western engineering universities, and the 7th Region I.R.E. Annual Achievement Award will be presented to the I.R.E. member in the Pacific Region adjudged to have contributed the most to electronics in the West during the past year.

Summer Session on Transistors Offered

Transistors and Their Applications will be the subject of a special summer program offered at the Massachusetts Institute of Technology from July 19 through July 30 during the 1954 Summer Session. The purpose of the program will be to outline the advantages and limitations of several important types of transistors in the context of their typical circuit applications.

Laboratory sessions demonstrating both physical principles and circuit applications of the transistor will supplement each lecture topic, according to Dr. Richard B. Adler, Assistant Professor of Electrical Communications, who will direct the program. Professor Adler will be assisted in presenting this program on transistors by members of M.I.T.'s research staff and by special lecturers.

Tuition for this two-week program will be $160. Academic credit will not be offered. Full details and applications may be obtained from the Summer Session Office, Room 7-103, M.I.T., Cambridge, Mass.

1953 National Telemetering Conference Record

The complete papers and talks, with illustrations, as presented at the 1953 National Telemetering Conference held in Chicago, Ill., are now available in bound form. The 230-page book, entitled "1953 National Telemetering Conference Record," contains 34 papers on Missile and Utilities Telemetering, and presents a large amount of up-to-date material on both new fields of investigation and applications data.

Copies of the Telemetering Record may be obtained from: The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y. The price is $2.00 per copy, postpaid in the United States.

Calendar of COMING EVENTS

IRE-WCEMA Western Electronic Show & Convention, Pan Pacific Auditorium, Los Angeles, Calif., August 25–27

IRE-AIEE-URSI Symposium on Information Theory, Massachusetts Institute of Technology, Cambridge, Mass., September 15–17

Cedar Rapids Conference on Communications, Cedar Rapids, Iowa, September 17–18

IRE Professional Group of Vehicular Communications Meeting, Rice Hotel, Houston, Texas, September 30–October 1

National Electronics Conference, Hotel Sherman, Chicago, Ill., October 4–6

IRE Professional Group on Nuclear Science Annual Conference, Sherman Hotel, Chicago, Ill., October 6–7

Symposium on Marine Communication and Navigation, Hotel Somerset, Boston, Mass., October 13–15

IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., October 18–20

IRE Baltimore Section PGANE East Coast Conference on Airborne and Navigational Electronics, Sheraton-Belvedere Hotel, Baltimore, Md., November 4–5

IRE-PIB Microwave Symposium, Engineering Societies Auditorium, New York, N. Y., November 8–10

IRE-AIEE Conference on Electronic Instrumentation and Nucleonics in Medicine, Morrison Hotel, Chicago, Ill., November 10–11

IRE Quality Control Symposium, Statler Hotel, New York, N. Y., November 12–13

Symposium on Fluctuation Phenomena in Microwave Sources, Western Union Auditorium, New York, N. Y., November 18–19

IRE Kansas City Section Annual Electronics Conference, Hotel President, Kansas City, Mo., November 18–19

IRE-AIEE-ACM Eastern Computer Conference, Bellevue-Stratford Hotel, Philadelphia, Pa., December 8–10

IRE-AIEE-NBS-URSI Conference on High Frequency Measurements, Hotel Statler, Washington, D. C., January 17–19

AIEE Winter General Meeting, Hotel Statler, New York, N. Y., January 31–February 4

1955 Southwestern IRE Conference and Electronics Show, Baker Hotel, Dallas, Tex., February 10–12

IRE National Convention, Waldorf-Astoria Hotel and Kingsbridge Armory, New York, N. Y., March 21–24

IRE Headquarters Adds New Building

The building shown above at the right was recently purchased by the IRE to provide much needed additional space for its headquarters staff and committees. It conveniently adjoins the present headquarters (left) and will be ready for occupancy this month.

IRE Expands Its National Headquarters

The IRE has added substantially to its national headquarters facilities by purchasing a six-story building at 5 East 79 Street in New York City, adjoining the main headquarters building at 1 East 79 Street. It is an architectural gem, completely fireproof, which the IRE has modernized, at the same time preserving the atmosphere of a residence which was featured in the *Architectural Record* as an outstanding example of a large city house.

The tremendous growth in IRE membership and activities during the past few years has made necessary additional office and meeting space. Since 1946 when the IRE took occupancy of the main building, membership has jumped from 18,000 to over 38,000. During this period 38 Sections have swelled to 73 Sections and 17 Subsections, Student Branches have been founded in 120 colleges, and 22 Professional Groups totaling 25,000 members have been organized. To keep pace with this large growth, the headquarters staff has more than doubled with the consequence that the main building has become overcrowded.

The demand for committee meeting rooms has also greatly increased and now exceeds the supply, with 369 meetings held there during last year.

The newly-acquired building, which will be occupied the latter part of this month, provides four additional meeting rooms and four floors of office space. The first and second floors will comprise meeting and reception rooms. The third floor will be occupied by the accounting department. The technical secretary and his staff will have the fourth floor, where an additional meeting room has been provided for technical committees. The fifth and sixth floors will accommodate the editorial department. The mailing address for these departments will still be 1 East 79 Street.

The main building will continue to house the executive secretary and his assistants, the office manager, and the departments dealing with general membership matters and services.

Purchase of the new building was made possible by the funds remaining in the Building Fund, which was raised during the War in order to provide the IRE with a headquarters building.

ECPD Survey on Awarding Professional Degrees

The Recognition Committee of the Engineers Council for Professional Development, under the chairmanship of R. H. Barclay has completed a survey of the awarding of the professional degree by various engineering institutions. The survey, with its excellent response—of the 146 institutions sent questionnaires, 142 responded—will serve as a basis for formulating recommendations concerning the practice of awarding the professional degree as a means of professional recognition.

Of the engineering schools surveyed, 86 award the professional degree while 62 do not. Of the 86 awarding the degree, 8 require resident graduate study and 4 include both professional experience and resident graduate study as prerequisites for awarding the professional degree.

In regard to their future plans concerning the professional degree, 69 colleges will continue awarding the degree, 2 will institute the professional degree, 13 will abandon it, 49 will continue not to award it, and 17 are uncertain as far as future plans are concerned. The survey shows that approximately $\frac{1}{8}$ of the schools offering the professional degree have either dropped it or are making plans to do so.

The Committee also reported an increase of approximately 80 per cent in the awarding of professional degrees in the last five years as compared with the previous five-year period. In the last 10 years, 1387 to 1398 professional degrees have been awarded, and of these, 917 to 922 have been awarded during the last five years.

The Committee believes this survey to be the most comprehensive of its type ever undertaken and will use it in formulating recommendations on the practice which they believe should be followed in respect to awarding the professional degree as a means of professional recognition.

"Contributors" and "IRE People" Shifted

Beginning with this issue, "IRE People" and "Contributors" will be found in the Advertising Section of the Proceedings. Page numbers will be given in the Table of Contents.

The following Contributors were inadvertently omitted from the June issue of the Proceedings. They are printed here in full.

❖

D. D. King

D. D. King (M'46) was born on August 7, 1919, in Rochester, New York. He received the A.B. degree in engineering sciences from Harvard College in 1942 and the Ph.D. degree in physics from Harvard University in 1946. He was a teaching fellow in physics and communication engineering in 1943, serving as a staff member of the preradar Officer's Training School at Cruft Laboratory, Harvard University. During 1945 he was a research associate at Cruft Laboratory. In 1946 he was appointed research fellow in electronics and in 1947 assistant professor of applied physics in Harvard University.

In 1948 Dr. King was appointed associate professor of physics in the Institute for Co-operative Research of the Johns Hopkins University, and in 1950 assistant director of the Radiation Laboratory.

Dr. King is a member of Sigma Xi and the American Physical Society.

❖

H. Scharfman

Howard Scharfman (S'47–A'50) was born in New York, N. Y. on December 27, 1924. He entered the U. S. Signal Corps in 1943 where he worked on radar and pulse modulation equipment. He left the service in 1946. In 1947 he received the BSEE degree from New York University, and in 1948 he was awarded the MSEE from Northwestern University.

After working at the Kew Gardens, Long Island, Sylvania Electric Company plant over the summer of 1948, Mr. Scharfman joined the teaching staff of the Polytechnic Institute of Brooklyn. He taught undergraduate electrical engineering and continued his graduate studies there until 1950. He then joined the Antenna Section of the Seattle division of Boeing Airplane Company where he did research and development on receivers, antenna pattern ranges, and aircraft antennas.

Since 1951, Mr. Scharfman has been a Research Associate at the Radiation Laboratory of The Johns Hopkins University, Institute for Co-operate Research. He has been working on VHF and UHF circuits and microwave scattering problems while completing graduate studies for the DSEE. He is a member of Eta Kappa Nu and Sigma Xi.

Professional Group News

Ultrasonics Engineering

The Professional Group on Ultrasonics Engineering held its third administrative committee meeting on March 25 in New York City.

Plans were made to sponsor a session on ultrasonics at the IRE Western Show and Convention (WESCON) which will be held in Los Angeles on August 25–27. These plans were made following a recommendation by the administrative committee to sponsor sessions on ultrasonics at the National Electronics Conference (Nec) and WESCON on alternate years in addition to the regular sessions of the IRE Convention every year. It was felt tht this arrangement would result in better geographic participation in the Group activities.

The PGUE papers to be presented at WESCON are: (1) "Composite Piezoelectric Resonators" by W. G. Cady, California Institute of Technology; (2) "Ultrasonic Cleaning of Miniature Devices" by Q. C. McKenna, McKenna Laboratories; (3) "Power Measurements in Ultrasonics" by O. E. Mattiat, Clevite-Brush Development Co.

Dr. W. G. Cady, former President of the IRE, is preparing a tutorial paper on Ultrasonics for the IRE Student Quarterly, a magazine to be distributed to all student members. It will discuss the nature and uses of ultrasonics, opportunities existing in this field, and those college courses which are recommended for workers specializing in this field.

Julius Bernstein, of the Edo Corporation, has been elected to the Administrative Committee position vacated by J. L. Hunter.

Nuclear Science Group Holds First National Annual Meeting

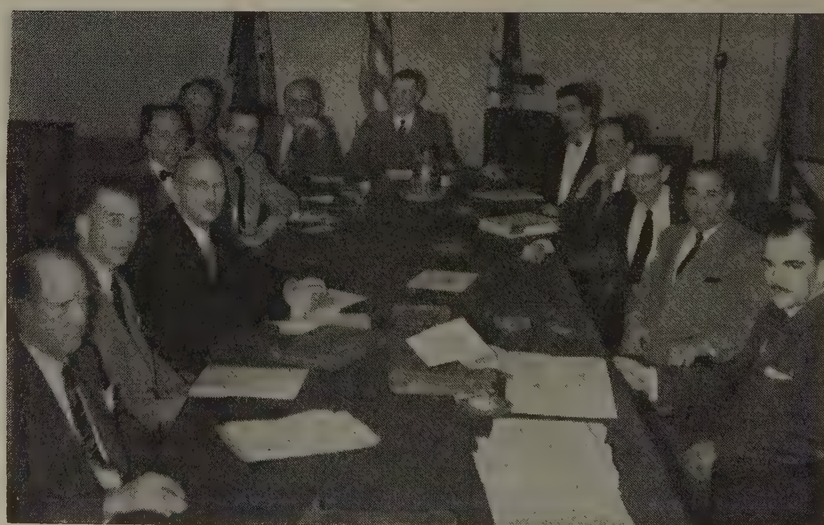
The Professional Group on Nuclear Science will hold its First National Annual Meeting in Chicago, Ill., on October 6 and 7, 1954, at the Sherman Hotel. The first day's session will coincide with the last day of the National Electronics Conference.

The program will be divided into the following four technical sessions of invited and contributed papers: (1) Nuclear Medical Electronics; (2) General Nucleonic Instrumentation; (3) European and American Reactor Technology; (4) European and American Particle Accelerator Technology.

Several noted European scientists will be invited to review their latest developments and a similar review of American developments will be presented. There will also be a nontechnical program Wednesday evening, October 6. It will consist of a cocktail party followed by motion pictures of nuclear weapons' tests and other significant documentary films.

Registration fees for this meeting will be as follows: Members of the Professional Group on Nuclear Science, $2.00; Members of the IRE, $3.00; Nonmembers, $4.00. Tickets to the cocktail party are $2.50 each. Members of the Professional Group on Nuclear Science who have paid their dues by September 1 will not only be admitted for $2.00, but they will also receive a copy of the PGNS Transactions.

Quality Control Symposium Committee



(*Left to right*)—E. Reardon, J. E. Gottschall, J. R. Steen, August Mundel, Leo Jacobson, P. Bobich, Irving Azoff, A. S. Marthens, E. J. Nucci, M. E. King, R. H. Cady, Leon Bass, and H. E. Herrick.

Quality Control Symposium

A National Symposium on quality control of electronic components is scheduled to be held in New York City on Friday and Saturday, November 12–13, at the Hotel Statler. This symposium is being sponsored jointly by the Professional Group on Quality Control of the I.R.E., and the Electronics Technical Committee of the American Society for Quality Control.

This two-day symposium, on a national scale, will have four sessions, featuring experts in the field of quality control from all branches of industry, research, and government agencies.

Some of the main topics to be covered are: (1) case histories of quality control applications in radio and television; (2) the relationship of military electronic equipment to quality control; (3) clinic of current quality control application problems in the electronic industry; (4) tutorial theory and philosophy of quality control.

Color Television Symposium

A Color Television Symposium, sponsored by the Philadelphia Section of the IRE, has recently been concluded. The six meetings of the symposium were held in the University Museum Auditorium on the campus of the University of Pennsylvania. An integrated coverage of technical information pertaining to color television was presented by speakers who were carefully selected from outstanding experts on each topic. Attendance was good at all meetings and the lectures were received enthusiastically.

In addition to six meetings held from February 11 to April 1, registrants were invited to a conducted tour of the NBC Television Studio at Colonial Theatre, New York City, on March 23. This tour was purposely arranged to fall during the week of the IRE National Convention. The Colonial Theatre tour was well attended and proved to be well worthwhile. Camera and control equipment was demonstrated in action with a live model. A complete color camera was arranged for the display of internal parts and competent guides were available to explain the function of the equipment and to answer questions. The lighting control room was also opened to visitors.

Radio Progress Report Discontinued by Board

On the recommendation of the Editorial Board, the IRE Board of Directors voted to discontinue the annual Report of Radio Progress, which appears each year in the April issue of Proceedings. The Annual Review Committee, which prepares the report, was discontinued also.

Other plans are now underway for providing Proceedings readers with informative reviews of recent progress in various branches of the radio engineering field. Details will be announced at a later date.

Northwest Florida Subsection Holds First Technical Meeting

The IRE held the first technical meeting of the newly-formed Northwest Florida Subsection on April 17, 1954, in the Air Force Armament Center's new Andrews Engineering Building at Eglin Air Force Base. Brigadier General Edward P. Mechling welcomed the new organization to the area, and stressed the importance to the Air Force of encouraging and co-operating with the IRE as well as other national engineering organizations. Donald B. Houghton, of the Franklin Institute in Philadelphia, was also a guest speaker at the meeting. The organization is a subsection of the Atlantic Section, and was approved by the IRE Executive Committee on March 2, 1954. Group membership has increased from 32 to 47 in its first month, indicating the need for the subsection in this area.

Abstracts of Transactions of the I.R.E.

The following issues of *Transactions* have just been published, and are now available from The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Antennas and Propagation	Vol. AP-2, No. 3	1.50	2.25	4.50
Audio	Vol. AU-2, No. 3	.95	1.40	2.85
Circuit Theory	Vol. CT-1, No. 2	1.00	1.50	3.00
Electronic Computers	Vol. EC-3, No. 2	1.65	2.45	4.95
Ultrasonics Engineering	PGUE-1	1.55	2.30	4.65

*** Public libraries and colleges can purchase copies at IRE Member rates.**

ANTENNAS AND PROPAGATION

Vol. AP-2, No. 3, July, 1954

Low Frequency Waves on Transmissions Lines of Composite Section—R. W. Klopfenstein

Three types of transmission lines are analyzed, and it is found that the principal wave is not transverse electromagnetic in transmission lines for which the medium between the conducting boundaries is other than a homogeneous isotropic dielectric. Calculations of characteristic impedance and of propagation on the basis of electrostatic and magnetostatic field distributions are accurate in the low-frequency limit. The cut-off frequency for the first higher order modes lies between those that would be obtained if the transmission lines were entirely filled with one dielectric material or the other.

Paraboloid Reflector and Hyperboloid Lens Antennas—E. M. T. Jones

A theoretical analysis of the radiating properties of the paraboloid reflector and the hyperboloid lens, shows that low amplitude cross-polarized radiation and high gain factors can be obtained from a paraboloid reflector excited by a plane-wave source. Low amplitude, cross-polarized radiation can also be obtained from the hyperboloid with a plane-wave feed, but with a lower gain factor. It is found that the measured properties of the antennas agree reasonably well with the theoretical predictions Also it is found experimentally that principal plane side lobes of the order of -40 db can be obtained with a short focal length hyperboloid lens.

Arrays of Closely-Spaced Nonresonant Slots—R. J. Stegen and R. H. Reed

Slots laid broadside to each other exhibit mutual coupling of such magnitude that the design of practical linear arrays of such slots has hitherto been difficult, if not impossible. The technique presented here will produce arrays capable of generating pencil beams or shaped with controllable side-lobe level. A large number of slots per wavelength are used and the mutual-coupling effects are kept small by making the slots short compared to a half wavelength. Tests have shown that these arrays may be placed side by side without interaction, thus making it possible to construct a two-dimensional array.

A Theoretical and Experimental Study of the Recombination Coefficient in the Lower Ionosphere—A. P. Mitra and R. E. Jones

The problem of recombination of electrons and ions in the lower ionosphere is studied both experimentally and theoretically. The experimental study involves analysis of new experimental data such as 150-kc radio-wave absorption polarization and phase heights; absorption of short wave galactic radiation; and E-region critical frequency, as well as recombination values already published. Then, by use of the theories of dissociative recombination and negative ions, a theoretical model is derived which is consistent with the experimental results. The values of the coefficient during night-time and during sudden ionospheric disturbances are discussed.

A New Antenna Feed Having Equal E- and H-Plane Patterns—Alvin Chlavin

When two complementary sources are combined in the proper amplitude and phase, desirable radiation characteristics for feeding a circular aperture are obtained. It is shown that when the feed is achieved there results a circular beam cross section which optimizes the efficiency of illumination of a circular aperture. The back radiation from the feed is down 30 db from that in the forward direction, minimizing interference effects between feed and aperture. It is the purpose of this thesis to show how a feed composed of complementary sources has been physically realized and to present and discuss experimental radiation and impedance data.

It is well known that the radiation pattern of an electric dipole is a circle in the H plane and a figure 8 in the E plane. An open-ended coaxial line carrying the TE_{11} mode is similar to magnetic dipole; i.e., the E plane is nearly circular while the H plane is like a figure 8. These two sources have been combined to produce a feed whose E- and H-plane patterns are of equal width.

The complementary source idea has been applied to feeds of both linear and circular polarization. The linearly polarized feed is excited from rectangular waveguide and is simple to fabricate. It can be easily matched over a broadband. This feed has been used to illuminate a 20-parabola with the result that the secondary E and H planes are of equal width and the side lobes are 30 db down from the main radiation. The circularly polarized feed is excited from a circularly polarized TE_{11} mode in coaxial line. The radiating structure maintains circular symmetry and the axial ratio remains essentially constant over a large portion of the beam.

Virtual Source Luneberg Lenses—G. D. M. Peeler, K. S. Kelleher, and H. P. Coleman

The portion of a spherical Luneberg lens contained between two plane reflectors has

been investigated as a lens of reduced size and weight. If the reflectors pass through the center of the sphere, the resulting system produces several perfectly focused radiation beams, each appearing to originate from a virtual source on the surface of the full sphere. The virtual source positions and the position, beamwidth, and gain of the beams are accurately predicted from the spherical wedge angle and the source position. When the wedge angle is π/p where p is an integer, rays with p reflections form the beam having the greatest gain and a displacement from the wedge bisector equal to the source displacement. For applications in which only this principal beam is desired, the gain of the unwanted beams can be reduced by absorption, reflection, or illumination taper.

Scanning is achieved by moving the feed along the surface of the spherical wedge; if p is an odd integer, scanning can be obtained by moving the wedge past a fixed feed.

Experimental data were taken on a two-dimensional X-band model having a value of $p=1$. Good agreement was found with the predicted performance regarding beam position, beamwidth, and gain. The single, undesired beam was minimized by the use of absorbing material.

AUDIO

VOL. AU-2, No. 3, MAY–JUNE, 1954

PGA News

PGA Chapter News

IRE Technical Committee News

Visible Speech—Rotary Field Co-ordinate-Conversion Analyzer—Friedrich Vilbig

A process is discussed, in which a rotary field is generated by the speech frequency spectrum. The rotary field deflects the electron current of an oscilloscope, so that typical polar co-ordinate pictures appear on the screen. These are substantially independent of fundamental frequency. Successive pictures can be produced by conversion of these polar co-ordinate pictures into cartesian co-ordinate pictures. In order to accomplish this, one employs a disc with a radial slit, rotating at a speed ν, and a glass prism, rotating at a speed of $\nu/2$. Selected parts of the screen picture can be traced through regulation of the intensity, e.g. by a selected range of the frequency spectrum. The cartesian pictures are thus also simplified, so that a "reading" of speech symbols becomes possible. In addition to its use for the speech analysis, this process may also be employed to produce polar and cartesian pictures of brain waves, etc. Moreover, polar co-ordinate maps can be converted into cartesian co-ordinates and vice-versa.

Dynamic Amplifiers for Phonographic Reproduction—E. S. Purington

In dynamic amplifiers for phonographs, the gain versus frequency characteristics change in accordance with the varying nature of the impressed audio signal. This paper outlines the basic principles of design of such amplifiers for expansion and for background noise reduction purposes. Especial reference is made to the fundamental patent literature in this field.

Components and Mechanical Considerations for Magnetic Sound on 35-MM Film—John F. Frayne

The art of sound recording in the motion picture field had reached a high professional status before magnetic recording began to receive consideration by the radio and television industries. This article reviews this status and discusses the relative merits of 35-mm sprocket-hole film and ¼-inch tape with respect to the requirements of the motion picture industry. The article traces the applications of the magnetic recording medium to single-track, multi-track and stereophonic recording, including composite magnetic sound and picture film. The discussion includes a description of the associated apparatus and its conformity to the established practices of the industry.

A Loudspeaker Accessory for the Production of Reverberant Sound—D. W. Martin and A. F. Knoblaugh

Organ music produced in small rooms having little natural reverberation, can be enhanced by the addition of artificial reverberation. A direct method of adding the aftersound at the electroacoustic transducer itself is described. This is in contrast to the more conventional reverberation systems employing driving transducers, time-delay means, pickup transducers, and mixing and amplifying circuits. Multiple resonant helical mechanical delay lines store the energy and radiate it at a later time. In one model the coupling to the transducer is mechanical, and in the other model acoustical coupling provides a number of practical and acoustical advantages.

CIRCUIT THEORY

VOL. CT-1, No. 2, JUNE, 1954

Simultaneous Oscillations in Oscillators—Johannes S. Schaffner

An oscillator with two degrees of freedom can, under certain conditions, oscillate simultaneously at two different frequencies. The ratio of the two frequencies may be rational (synchronous oscillations) or irrational (asynchronous oscillations). The conditions necessary for simultaneous oscillations are discussed for these two cases. The problem is essentially nonlinear and cannot be solved by the methods of linear analysis. The method of equivalent linearization, developed by Kryloff and Bogolinkoff, is shown to give a good correlation with experimental results.

Synthesis of Transfer Functions by Active RC Networks—D. B. Armstrong and F. M. Reza

The paper considers the synthesis of arbitrary passive transfer functions using stable active RC networks.

The work undertaken is oriented toward discussing general methods of synthesis rather than specific design procedures. The structures used consist of single unidirectional feed-back loops. These methods, although by no means providing the most general procedure for the synthesis of active networks, appear to offer a logical step toward this aim.

Reciprocity Relations in Active Three-Terminal Elements—Jacob Shekel

A vacuum-tube of a transistor, operating linearly under small-signal conditions, may be represented as a resistive (i.e., nonreactive), three-terminal network element. Such an element may be considered as composed of a triangle of resistors and a three-terminal gyrator. If the element is active, at least one of the resistors is negative.

The activity or passivity of the element is shown to be dependent on the three resistors, while the reciprocity relation is obeyed or violated according to whether the gyrator is absent or present, respectively. It is also shown that an active three-terminal element cannot obey reciprocity and remain stable, so that the gyrator has a stabilizing effect.

Some general relations are derived concerning three-terminal active elements. These relations apply to any active element that is described by a three-terminal resistive network, without regard to the physical principles underlying its operation.

Power Gain in Feedback Amplifiers—Sam J. Mason

A linear three-terminal device Z is imbedded in a lossless passive network N and the properties of the complete system, as measured at two specified terminal pairs, are described by the open-circuit impedances Z_{11}, Z_{12}, Z_{21}, Z_{22}. A search for properties of Z which are invariant under the transformation N leads to the quantity

$$U = \frac{|Z_{21} - Z_{12}|^2}{4(R_{11}R_{22} - R_{12}R_{21})},$$

where R_{jk} is the real part of Z_{jk}. Quantity U is independent of the choice of N and is (consequently) invariant under permutations of the three terminals and also under the replacement of the open-circuit impedances by short-circuit admittances. If U exceeds unity at a specified frequency, then N can always be chosen to make R_{11} and R_{22} positive and Z_{12} zero at that frequency. Quantity U is identifiable as the available power gain of the resulting unilateral structure.

An arbitrary coupling network may be decomposed into a portion which accomplishes unilateralization and a remaining complementary portion which provides feedback around the unilateralized structure. Such decomposition brings some of the notions of elementary feedback theory to bear upon nonunilateral circuit analysis and offers a viewpoint from which signal flow and power flow are simply related.

Spectra of Waves with Periodic Modulation—Leo V. Skinner

The harmonic analysis of a wave with amplitude $a(t)$ and angle modulating function $\psi(t)$ is simplified for cases in which $a(t)$ and $\psi(t)$ are periodic and their periods are rationally related. Further simplifications are made when these functions have symmetrical properties. Examples are given to illustrate the procedure.

What is Nature's Error Criterion?—Maurice V. Joyce (*correspondence*)

Network Synthesis and the Moment Problem—Roy C. Spencer (*correspondence*)

The Champions—H. J. Carlin (*correspondence*)

PGCT News

ELECTRONIC COMPUTERS

VOL. EC-3, No. 2, JUNE, 1954

PGEC News

Logic, Discovery, and the Foundations of Computing Machinery—M. E. Maron

This paper describes the logical nature of computing machines in terms of languages and the types of problems that can be solved by logical operations on languages. The problem of *discovery* in mathematics and empirical science is discussed, and an "inductive" machine is described which would be able to formulate hypotheses, modify them in the light of new experience, and eventually discover the laws of a very simple universe.

System Design of the SEAC and DYSEAC—A. L. Leiner, W. A. Notz, J. L. Smith and A. Weinberger

In the course of developing the system plans for the DYSEAC and SEAC, certain standard methods and procedures were evolved for producing a large-scale digital computer design. These standard procedures cover, first, the development of system specifications, second, the development of functional plans, and finally, the development of wiring plans. The later stages of these procedures are reducible to se-

quences of simple steps, capable of being systematically formulated in explicit terms. The similarity between these procedures and many of the data-processing procedures commonly being executed by present-day computers suggests that, with further development of these design techniques, the wiring plans for new computer systems might well be produced by existing digital machines.

Digital Techniques in Analog Systems—M. A. Meyer

This paper discusses analog computation where the analog components consist of digital elements. Pulse rate is the quantity used to represent the data. Such computation systems may have many advantages over present analog techniques. The various components necessary to produce a complete computation system are described. Several examples of the use of these components to solve specific problems are shown.

A High Speed Correlator—Harold Bell, Jr. and Vincent C. Rideout

The correlation function,

$$\phi_{ij}(\tau) = \lim_{-T\to\infty} \frac{1}{T}\int_0^T f_i(t)f_j(t-\tau)dt$$

is of great interest today because of its use in the fields of oceanography and meteorology and because of its recent applications in the field of communication. Various machines, both analog and digital, have been designed for the automatic computation of correlation functions. The machine described in this paper differs from those which have previously been described in the literature in that the speed with which it computes the integral above for each value of τ_i is mainly limited by the minimum value of T permissible for a precision of a few per cent.

The high-speed correlator is of the analog type and utilizes the following main elements:

1. A high-speed (or wide-band) multiplier.
2. A high-speed integrator.
3. A high-quality lumped-constant delay line with a total delay of 2,080 microseconds.
4. A switching scheme which permits the delay τ_i to be varied rapidly in discrete steps.

With these units the correlator developed allows a correlation function of 41 discrete points to be computed for functions occupying nearly the complete audio spectrum, in a period of only five seconds.

A number of tests of this machine operating as an auto-correlator and as a cross-correlator have been made. Theoretical studies regarding precision have been checked by such tests for simple functions, and show that a precision of 2 per cent is possible. Separation of a sine-wave signal from noise was found possible for S/N ratios of as low as -25 db.

In addition to other possible uses, the high speed of this correlator has raised the possibility of using it to continuously process data in cases where, for example, a signal is masked by noise. With the aid of two tape recorders a single high-speed correlator should make it possible to continuously process the data in a band 360 cycles wide, with a two-second delay. Signals occupying wider bands would require more correlators for continuous processing, but could be handled if they are known to occupy the band for only certain calculable percentages of the time.

A Wide-Band Square-Law Computing Amplifier—Aaron S. Soltes

Wide-band computing amplifiers capable of accurately and continuously yielding output signals with amplitudes proportional to the mathematical square of their input amplitudes have been developed. These computers are designed to be inserted in signal channels in much the same manner as conventional amplifiers, but impart thereto a square-law transfer characteristic. Details are given concerning a model that provides an accuracy of the order of 1 per cent of full scale over an output dynamic range of 40 db when operating on fractional microsecond, pulsed carrier signals.

An Analog Multiplier Using Thyrite—L. D. Kovach and W. Comley

An investigation into the use of Thyrite as an inexpensive nonlinear element in the electronic analog computer indicates that this material has great value as a device capable of delivering an output voltage proportional to the square of the input voltage. The factors discussed are the characteristics of the material and the means by which these may be modified to produce a device capable of squaring with an accuracy of 1.25 per cent from dc to frequencies in excess of 1,000 cps. This ability to square makes possible the more important operation of the multiplication of two variable voltages.

A Sub-Audio Time Delay Circuit—C. D. Morrill

Through the use of an electronic differential analyzer arranged to give a sixth-order approximation of the Laplace shift operator, it is possible to reproduce an input signal and delay it T seconds when the highest angular frequency present in the input signal does not exceed $12/T$. Inaccuracy due to nonlinear phase shift is less than 2 per cent under these conditions. This device requires only linear computing elements, and will permit delays of seconds within a finite frequency spectrum. An application to a closed loop industrial control problem is cited and other applications are suggested.

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Review Section

Institutional Listings

ULTRASONICS ENGINEERING

PGUE-1, JUNE, 1954

History, Plans and Policies of the PGUE—A. L. Lane

On Membership—Who We Are and Where—M. D. Fagen

Ultrasound and Medicine—Julia F. Herick and Frank F. Krusen

Many therapeutic procedures have been introduced into the art and science of medical practice empirically, by a method of trial and error. Such a method has marked limitations and may produce unfavorable consequences. The facilities available in experimental medicine permit a thorough and extensive investigation of biological effects of many therapeutic agents or procedures before any clinical application is made. Optimal doses of the therapeutic agent can be approximated and a comparative study of various techniques designed for clinical application can be made. Experimental measurements have shown clearly that the major effect of ultrasound on living tissues is thermal when an intensity suitable for therapy is employed. The heating of bone by ultrasound is spectacular. No other agent used in therapy can raise the temperature of bone to such high levels so quickly. The application of ultrasound for diagnostic purposes and also for instrumentation will be described.

A Noncontact Micro-Displacement Meter—Harold F. Sharaf

A method capable of detecting micro-inch displacements of a transducer surface vibrating in the fractional megacycle range has been developed. The dynamic displacements of a vibrating transducer are detected with the aid of a specialized frequency-modulation system. The vibrating transducer face is incorporated into the system in a manner such that the center frequency of an oscillator is varied in the accordance with the transducer vibrations. The frequency-modulation system transforms the frequency deviations into a corresponding amplitude varying signal. In this way, accurate measurements of transducer vibrations can readily be obtained. The system is inherently capable of detecting transducer displacements with equal facility in either air or a viscous fluid.

Characteristics of Ultrasonic Delay Lines Using Quartz and Barium Titanate Ceramic Transducers—John E. May

Previous analysis of the equivalent circuit for the ultrasonic delay line was applied to quartz crystal transducers and a vitreous silica delay medium. The analysis has been extended to evaluate the use of other material such as magnesium, aluminum, and steel, and to include $BaTiO_3$ ceramic transducers. Calculations show the decrease in loss to be expected as the impedance of the delay medium is decreased and, conversely, the increase in bandwidth to be expected as the impedance of the delay medium is increased. The increase in bandwidth resulting from symmetrical loading of the transducers is also evaluated. Modifications to the equivalent circuit for the case of $BaTiO_3$ ceramic transducers are considered. Theoretically, losses can be reduced to 10 db or less using the ceramic transducers with a vitreous silica medium. The bandwidth is slightly less than for quartz transducers. However, by utilizing higher impedance materials, bandwidth approaching 100 per cent should be achieved.

Metal Cleaning and Its Improvement by the Use of Ultrasonics—T. J. Kearney

Ultrasonics is one of the newest tools to be applied to metal cleaning. Through the years abrasives, soaps, compounded alkaline materials, and solvents have been used to remove soils from metals. Trichlorethylene Solvent degreasing has had a wide reception in industry since its introduction in 1930. The Detrex Soniclean process combines sound energy and trichlorethylene solvent degreasing for metal cleaning. This is accomplished by immersing especially treated barium titanate transducers in chlorinated solvents producing cleaning results previously unobtainable. Crossrod, conveyorized equipment incorporating constant distillation and filtration of the solvent in the sonics chamber and providing a final vapor rinse and drying is now producing as many as 8,000 parts per hour in industrial plants.

A Temperature-Controlled Ultrasonic Solid Acoustic Delay Line—E. A. Pennell (*abstract only*)

An Acoustic Flowmeter Using Electronic Switching—H. P. Kalmus, A. L. Hedrich, and D. R. Pardue

A method of measuring fluid flow velocity is described wherein the difference in acoustic delay times in the upstream and downstream directions are compared to determine the velocity. The functions of two transducers are interchanged between transmitting and receiving at a rate of 100 times per second. Synchronous rectification techniques are employed to remove the flow information. The method described utilizes an all-electronic method of switching transducer functions and an ultrasonic wave of 185 kilocycles. Features of the system include fast response times—of the order of 0.05 second; and high sensitivity—flows as low as 1 cm/second. A complete description of the circuitry is given.

Books

Magnetic Amplifiers by George M. Ettinger

Published (1953) by John Wiley & Sons, Inc., 440 Fourth Ave., New York, N. Y. 76 pages+4-page index+5-page bibliography+1-page appendix+viii pages. 48 figures. $6\frac{3}{8}\times 4\frac{1}{2}$. $1.50.

The twofold purpose of this book is clearly stated in the preface. First it is intended to help the practicing engineer or physicist select the most suitable device to perform a given function. Second it is intended to serve as a basis for the research worker aiming to make further advances in the field.

The first purpose is fairly well accomplished with a survey and description of magnetic amplifier circuits drawn from the literature of the past twenty years. This survey is supported by an extensive bibliography. Possibly due to the brevity of the book, the author confines his discussion of each circuit to a simplified physical description of its operation and leaves any mathematical analysis to the reader, aided where possible by the references listed in the bibliography. No discussion is given on the subjects of size, weight, cost and relative performance of different circuits although these subjects are pertinent to the practical problem of selecting suitable magnetic devices for any given function.

In several places statements are made which are, no doubt, intended to simplify but actually tend to be loose or even misleading. This shows up at the beginning of Chapter Two where, by neglecting the demagnetizing effect of a-c, one is led to within a short step of designing a magnetic rectifier. In Chapter Three rotation of the axes of a transfer characteristic is purported to predict the effect of feedback. Actually only one of the axes should be rotated to get the desired result.

The second purpose of the author seems too ambitious for a book of this length. Since the book is well organized and readable it might serve very well as an introduction to the general field of magnetic amplifiers for the student or for one not already familiar with the field of electrical controls, if used with due consideration for its limitations. The topics discussed are well chosen to provide a good cross-section of the magnetic amplifier field and cover the more recent developments as well as the earlier work.

In this text you are introduced to the magnetic amplifier by a discussion of the saturable reactor. You are then led to the application of feedback and on to the more complex types of amplifiers such as the harmonic type and the ferro-resonant type. Time constant and transient response are then considered. The final chapter considers the rapidly expanding application of saturating core elements to computers. At the end of the book the bibliography and a reading list are included along with a chart which compares some of the core materials now being used in magnetic amplifiers.

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Thermionic Valves by A. W. H. Beck

Published (1954) by Cambridge University Press, Bentley House, London N.W. 1, England and 32 East 57th Street, New York 22, N. Y. 524 pages+4-page index+28-page appendix+4 pages tables+7 pages recent references+xvi pages. 211 figures. $5\frac{3}{4}\times 8\frac{3}{4}$. $12.00.

A. W. H. Beck is with the Standard Telephones & Cables, Ltd., England.

Although this book bears the broad title "Thermionic Valves," it covers only high-vacuum valves with thermionic cathodes. It is divided into three main parts: (1) the physical theory of electronics; (2) the mathematical theory of electronics; and (3) types of valves. Actually the mathematical treatment of all three of these parts is almost equally intensive so the book will be of most interest and value to those who gain new knowledge or refresh their old with the aid of mathematical treatment.

In the part devoted to valve types, there are chapters or sections on such types as velocity modulation, ultrahigh-frequency triodes, traveling-wave valves, beam interaction valves, magnetrons, picture convertors and storage devices, as well as the more or less conventional transmitting and receiving types.

The text will be of particular interest to those engaged in valve design rather than application. Also, as the author states, the book is "strongly biased toward microwave valves." With few exceptions, there are no portions of the text devoted to specific types of either British or American valves and there is included little or no material on valve manufacture or mechanical design.

Although there are no reproductions of photographs included, there is a generous use of diagrams, drawings and graphs.

Throughout the text, many references are listed and included at the end of the book is a seven-page appendix which lists for each chapter the most recent references on the subject matter involved. There are also appendices reporting the most recent work on oxide cathodes, noise in traveling-wave valves and propagation of space-charge waves on electron beams resulting from the work of Hahn and Ramo.

It is impossible, of course, to treat exhaustively in one volume all of the subjects included in this book, but the generalized treatment is probably ample for those with a good foundation knowledge of this branch of science.

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Complex Variable Theory and Transform Calculus—Second Edition by N. W. McLachlan

Published (1953) by Cambridge University Press, Bentley House, N.W. 1, London, England, and 32 E. 57th Street, New York, N. Y. 334 pages+10-page index+40-page appendix+xi pages. 102 figures. $5\frac{1}{2}\times 8\frac{3}{4}$. $10.00.

N. W. McLachlan is Professor of Electrical Engineering at the University of Illinois, Urbana, Ill.

This book is divided into three parts. The first part (131 pages) is devoted to "Theory of Complex Variable." In the words of the author this part *"must be understood thoroughly."* The second part (60 pages) deals with "Theory of Transform Calculus." The last part (140 pages) is a collection of technical problems representing applications of the first two parts. The illustrative examples are taken from many fields: electric circuits and transmission lines, airplane dynamics, deflection of beams, radio receivers, television amplifiers, submarine cables, electrical wave filters, condenser microphones, loudspeakers, heat problems connected with refrigerators, automobile brakes, annealing of steel rods. Illustrations and examples are sprinkled liberally into the first two parts as well. The chief strength of the book is in this wealth of illustrative material. It is for this reason that the book will appeal to engineers.

As far as the basic theory is concerned, it appears that the author expects from his readers some prior knowledge and understanding of functions of a complex variable. Thus on page 8 the author illustrates the differentiation of functions of a complex variable by differentiating e^z without having previously explained the meaning of this function when z is imaginary or complex. Other transcendental functions, such as log z, sin z, sinh z, and so forth are not explained either. Apparently the author expects his readers to treat the complex variable as if it were the real variable and use the formulas which are normally derived for functions of the real variable. The author claims that "the chapters on Complex Variable have been . . . made rigorous enough for all but the pure mathematician, *to whom the book is not addressed.*" As far as rigor is concerned, this reviewer is willing to accept the author on his own terms; but more thorough and more careful explanations in the early stages of the development of complex variable theory might have greatly added to the value of the book.

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1954 Convention Record of the I.R.E.

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The Index to the Abstracts and References published in the PROC. I.R.E. from February 1953 through January 1954 is published by *Wireless Engineer* and included in the March 1954 issue of that journal. Copies of this issue may be purchased for $1 (including postage) from the Institute of Radio Engineers, 1 East 79th Street, New York 21, N.Y. As supplies are limited, the publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals canned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

A new section has been introduced covering the general technique of electromagnetic waves, oscillations and pulses (i.e. transmission lines and circuits), as distinct from specific applications to telecommunications. The new section is numbered 621.37, with subdivisions. Full details of the new classification, and of the numbers which become obsolete as a result of its introduction, are given in PE Note 535, obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution 2 Park Street, London, W.1., England.

Section 621.396.67, dealing with Antennas, has been modified and expanded; details of the new classifications are given in PE Note 519.

Section 621.396.96, with subdivisions, has been introduced to cover Radar; details of the new classifications are given in PE Note 518.

New Subject Section

A section headed Automatic Computers has been introduced.

ACOUSTICS AND AUDIO FREQUENCIES

534.121.1 **1631**

Concerning Combined and Degenerate Vibrations of Plates—M. D. Waller. (*Acustica*, vol. 3, pp. 370–374; 1953.)

534.2-13 **1632**

Combined Translational and Relaxational Dispersion of Sound in Gases—M. Greenspan. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 70–73; Jan., 1954.)

534.2-14 **1633**

Wave Propagation in a Randomly Inhomogeneous Medium: Part 2—D. Mintzer. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1107–1111; Nov., 1953.) The region of validity of the single-scattering approximation used in part 1 (931 of April) is found by considering the next higher approximation. Some results are given for the case where the sound wavelength is large compared with the size of the inhomogeneity.

534.213.4:534.833.1 **1634**

Transmission of Sound through a Stretched Membrane—U. Ingard. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 99–101; Jan., 1954.) Analysis is given for a system comprising a membrane stretched across a tube. Conditions at resonance and antiresonance are studied by means of equivalent electrical circuits. Measurements made with a tube of diameter 3 inches and a fundamental frequency of 300 cps gave results in agreement with the theory.

534.213.4-13 **1635**

The Attenuation of Sound in Small Tubes—G. T. Kemp and A. W. Nolle. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1083–1086; Nov., 1953.) Report of measurements over the frequency range 3.8–20 kc, using air-filled tubes of internal radius 0.238 and 0.0292 cm respectively. The results are compared with values calculated from the Kirchhoff theory.

534.231 **1636**

Mean Force on a Sphere in a Spherical Sound Field: Part 1 (Theoretical)—T. F. W. Embleton. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 40–45; Jan., 1954.) Theory developed by King (*Proc. Roy. Soc. A*, vol. 147, pp. 212–240; 1934.) is presented in a form suitable for calculating the radiation force on a sphere in the field of a progressive spherical wave. The analysis is valid for all values of the ratio of wavelength to sphere radius. For large distances from the source the force is a repulsion varying as the inverse square of the distance. As the source is approached the force becomes an attraction, at a point depending on the frequency and on the sphere radius.

534.231:534.6 **1637**

Mean Force on a Sphere in a Spherical Sound Field: Part 2 (Experimental)—T. F. W. Embleton. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 46–50; Jan., 1954.) Measurements were made of the deflection of hollow glass spheres, of radius 0.228–3.029 cm, suspended by fine glass fibres in the field of a progressive sound wave, using frequencies between 300 and 7000 cps. Results give values of the force on the sphere in good agreement with values derived from the theory in part 1 (1636 above).

534.232:534.321.9:538.652 **1638**

Investigations on Magnetostrictive Oscillators at Frequencies between 80 and 500 kc/s—G. Steinkamp. (*Acustica*, vol. 3, pp. 399–404; 1953.) Measurements of the electromechanical and the mechanoacoustic efficiencies of five oscillators were made, from which the electroacoustic efficiency was deduced. Hysteresis and eddy-current losses were also determined. The losses due to internal magnetization processes are discussed in detail.

534.321.9:534.22.094.1 **1639**

Modulation Method of Measurement of Ultrasonic Dispersion—V. A. Zverev. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 91, pp. 791–794; Aug., 1953. In Russian.) The method was used in the experimental determination of dispersion in a long Ni wire, using a 1-mc ultrasonic wave modulated at 100 kc. Fractional differences of velocity down to $\sim 2.4 \times 10^{-3}$ were readily observed.

534.321.9:534.23 **1640**

Scattering of Underwater Plane Ultrasonic Waves by Liquid Cylindrical Obstacles—S. J. Bezuszka. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1090–1095; Nov., 1953.) Report of a theoretical investigation.

534.321.9-14 **1641**

Ultrasonic Transmission through Porous Bodies in Liquids—G. Schmid and H. Knapp. (*Z. angew. Phys.*, vol. 5, pp. 463–472; Dec., 1953.) Report of measurements at 350 kc using, clay diaphragms in a n/1000 KCl solution.

534.373:534.213 **1642**

Boundary-Layer Absorption in a Spherical Resonator—I. D. Campbell. (*Acustica*, vol. 3, pp. 395–398; 1953.) Calculation of the effective wall admittances representing viscous and thermal losses at the boundary leads to a solution of the decay of symmetrical modes of oscillation of a gas within a closed spherical resonator. A numerical example is calculated which permits assessment of the relative amounts of the absorption due to viscosity and thermal conductivity.

534.373:534.414 **1643**

A Study of the Factors influencing the Damping of an Acoustical Cavity Resonator—R. F. Lambert. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1068–1083; Nov., 1953.) Absorption measurements were made using a transverse-particle-velocity pickup device [1057 of 1950 (Hartig and Lambert)]. The results indicate that mechanical losses due to wall vibrations are experimentally significant. Empirical formulas are derived for the relaxation time of air molecules.

534.414 **1644**

On the Theory and Design of Acoustic Resonators—U. Ingard. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1037–1061; Nov., 1953.) Absorption and scattering from resonators in a free field and in walls are discussed. The effect on the resonance frequency of different aperture geometries is investigated. Design charts are presented for obtaining maximum resonance absorption, taking account of viscosity, heat conduction, and radiation.

534.414:534.231 1645
The Near Field of a Helmholtz Resonator Exposed to a Plane Wave—U. Ingard. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1062–1067; Nov., 1953.) Calculated values for the near field of a spherical resonator, assuming a uniform velocity distribution in the aperture, are in good agreement with measured values. The case of a resonator at the end of a tube is also discussed.

534.614-14 1646
Measurement of the Velocity of Sound in the Ocean—R. K. Brown. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 64–67; Jan., 1954.) An arrangement providing a continuous record of sound velocity as a function of depth is described. A phase-comparison principle is used; the operating frequency is 500 kc.

534.614-14 1647
Velocity of Sound in Glycerol—F. A. A. Fergusson, E. W. Guptill and A. D. MacDonald. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 67–69; Jan., 1954.) A method for measuring the velocity of sound in liquids is based on the principle of a method used to determine critical angle in optics. The liquid is placed in a thin-walled cell which is rotated in a tank of a second liquid for which the velocity of sound is known. The value obtained for glycerol at 25 degrees C. is 1964 ±10 m at a frequency of 7.5 mc. Values for glycerol-water mixtures are also reported.

534.614-14 1648
A Method of Measuring the Velocity of Sound in a Few Grams of a Liquid—J. H. Janssen. (*Acustica*, vol. 3, pp. 391–394; 1953.) A slight modification of the em sound generator of St. Clair (2455 of 1941) permits the determination of the resonance frequency of a bar clamped at its center, with and without a liquid column on top of it. The velocity of sound in the liquid can then be calculated.

534.614-14:534.321.9 1649
The Measurement of the Velocity of Sound in Liquids—G. M. Graham. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1124–1127; Nov., 1953.) A technique suitable for use in the frequency range 0.5–5 mc is described. Measurements on distilled water are reported.

534.64 1650
The Acoustic Impedance of a Porous Layer at Oblique Incidence—J. S. Pyett. (*Acustica*, vol. 3, pp. 375–382; 1953.) The specific normal impedance of a layer of anisotropic porous material for a plane wave incident in any direction is calculated in terms of the two propagation parameters of the material, on the assumption that one of the principal axes of the material is normal to the surface. Experiments on the same lines as those of Shaw (302 and 303 of February) gave results for rockwool in agreement with calculations.

534.76 1651
Basic Principles of Stereophonic Sound—W. B. Snow. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 61, pp. 567–587; Nov., 1953. Discussion pp. 587–589.) A review of published theory, with examples of practical applications. 68 references.

534.79 1652
A Technique and a Scale for Loudness Measurement—W. R. Garner. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 73–88; Jan., 1954.) A true ratio scale of loudness is obtained by interrelating two scales derived from judgments made on two different principles, namely (*a*) equisection judgments, in which the observer adjusts a series of tones between two fixed limits to give a series of equal loudness intervals, and (*b*) fractionation judgments, in which the observer is instructed to adjust a tone to obtain a specified fraction of the initial loudness. Results of experiments indicate the validity of the assumptions made.

534.79 1653
Theory of Loudness Level and Loudness Sensation—G. Quietzsch. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 169–177; 1953.) A summarized account of recent work on loudness level and loudness sensation with particular reference to the relation between the phon and sone scales.

534.833 1654
Electronic Sound Absorber—H. F. Olson and E. G. May. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1130–1136; Nov., 1953.) A microphone, amplifier, and loudspeaker are arranged in a feedback system so as to reduce the sound pressure near the microphone. Reductions of 10–25 db can be achieved over a three-octave range at low audio frequencies. A suitable microphone is the electronic type presviously described [3413 of 1947 (Olson)].

534.833 1655
Measurement of Sound Absorption Coefficients at the Research Centre for Cinematography in Rome—G. Parolini. (*Ann. Télécommun.*, vol. 8, pp. 391–394; Dec., 1953.) Measurements were made with warble tones in a reverberation chamber. Results for 31 types of sound-absorbent materials of Italian manufacture, and for costumes of various periods, are tabulated.

534.834:534.861.1 1656
Acoustical Measurements on Components of Air-Conditioning Installations for Broadcasting and Television Studios—G. Venzke. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 224–231; 1953.)

534.84 1657
An Empirical Acoustic Criterion—T. Somerville. (*Acustica*, vol. 3, pp. 365–369; 1953.) A parameter R has been evolved which expresses the divergence of the reverberation-time/frequency characteristic from the mean reverberation time T_m. A parameter D has also been evolved which is a measure of the general decay irregularity for the studio under test. When $(D+0.7\ R)$ is plotted against T_m, the points representing studios and halls subjectively classified as having good acoustic properties are all found to lie within a particular area.

534.85 1658
Reduction of Background Noise in the Reproduction of Gramophone Disk Records—M. J. de Cadenet. (*Rev. Son*, no. 9, pp. 311–322; Dec., 1953.) A review of basic circuits for use particularly with modern microgroove records.

621.395.61 1659
An Acoustic Lens as a Directional Microphone—M. A. Clark. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1152–1153; Nov., 1953.) Discussion of the radiation pattern of a microphone in which the transducer is located at or near the focus of a lens, e.g. of slant-plate type, the two being joined by a conical horn.

621.395.623.7 1660
A Physical Approach to the Generalized Loudspeaker Problem—O. K. Mawardi. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1–14; Jan., 1954.) The problem formulated by Foldy and Primakoff (264 of 1946) is studied and an expression is derived relating the sound pressure at a point in the field to the electrical signal at the loudspeaker terminals, a transfer function being defined by analogy with the methods of network analysis. For the simple case of a circular loudspeaker mounted in an infinite baffle, the integral equation obtained is solved exactly.

621.395.625(083.74) 1661
I.R.E. Standards on Sound Recording and Reproducing: Methods for Determining Flutter Content, 1953—PROC. IRE, vol. 42, pp. 537–541; March, 1954.) Standard 1953 I.R.E. 19S2.

621.395.625.3 1662
Accurate Measurement of the Speed of Transport of Magnetic Tapes—F. Gallet. (*Rev. Son*, no. 7, pp. 252–255; Oct., 1953.) The mains voltage is recorded on a length of the tape under test, and the magnetic-spectrum method is applied to locate the positions of the alternations. The length of tape occupied by 100 alternations gives the speed that would be obtained with a constant main frequency of 50 cps. An alternative method for continuous measurements is mentioned.

621.395.625.6:537.228.3 1663
Electro-Optic Sound-on-Film Modulator—R. O'B. Carpenter. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1145–1148; Nov., 1953.) Description of a variable-density sound-recording system using a crystal modullator of the type described previously [2209 of 1953 (Dressler and Chesnes)].

621.395.92 1664
Recent Advances in Hearing Aids—L. G. Hector, H. A. Pearson, N. J. Dean, and R. W. Carlisle. (*Jour. Acoust. Soc. Amer.*, vol. 25, pp. 1189–1194; Nov., 1953.) Three models are described; the output powers are about 1, 5, and 40 mv respectively.

ANTENNAS AND TRANSMISSION LINES

621.372.2 1665
Experimental Study of the Transmission ot Centimetre Waves along Wire Waveguides—P. Chavance and B. Chiron. (*Ann. Télécommun.*, vol. 8, pp. 367–378; Nov., 1953.) Measurements over the frequency range 3–10 kmc are reported; six different polyethylene-covered Cu wires were used. Various types of launching device are described. Results indicate that the greater the reduction of phase velocity, the greater is the concentration of energy round the guide; the degree of concentration increases with the thickness of the dielectric coating. Discontinuities give rise to both reflection and radiation. Rain caused attenuations of 3–7 db /100 m at 3.15 kmc; a film of soot produced an attenuation of only 1.5 db/100 m. A power of 250 kw was transmitted without complications. Possible applications are discussed.

621.372.2 1666
Attenuation and Power-Handling Capability of Helical Radio-Frequency Lines—J. H. Bryant and E. J. White. (*Trans. I.R.E.*, vol. MTT-1, pp. 33–36; Nov., 1953.) A method is given for calculating the cold insertion loss and the power-handling capacity of a helical line with or without an outer coaxial cylindrical conductor.

621.372.2+535.41]:621.3.012 1667
On a Property of a Family of Equiangular Spirals and its Application to Some Problems of Wave Propagation—K. Landecker. (*Jour. Appl. Phys.*, vol. 25, pp. 41–48; Jan., 1954.) The locus of all end points of the normals drawn from a point in a plane to the members of a family of equiangular spirals is a circle. The standing-wave pattern in a uniformly attenuating medium may, in simple cases, be represented by a vector diagram using a symmetrical pair of spirals. Construction of the locus circle enables the voltage and current maxima and minima to be determined. Transmission-line and light-wave interference examples are given, and the application in conjunction with a Smith chart is indicated.

621.372.2:621.315.212:621.317.733 1668
Some Steady-State Characteristics of Short Irregular Lines—A. Rosen. (*Electronic Eng.*, vol. 26, pp. 90–96; March, 1954.) Analysis is based on representation of the distribution of impedance irregularities by a Fourier series. Measurements of impedance and of attenuation coefficient are discussed. A description is given of a bridge with variable-capacitance ratio arms for testing factory lengths of coaxial cable. See also 1295 of May.

621.372.21.029.6 1669
Low-Loss Microwave Line—F. J. Tischer. (*Arch. elekt. Übertragung*, vol. 7, pp. 592–596; Dec., 1953.) Analysis of wave propagation along a line consisting of parallel strip conductors and a dielectric cross-bar. The transverse magnetic field parallel to the conductors is zero, hence no induced electric currents flow along the strips. Sections can therefore be coupled directly. The attenuation/wavelength characteristics of the line are compared with those of rectangular and circular waveguides.

621.372.8 1670
Propagation of Microwaves through a Cylindrical Metallic Guide Filled Coaxially with Two Different Dielectrics: Part 2—S. K. Chatterjee. (*Jour. Indian Inst. Sci.*, Section B, vol. 35, pp. 103–117; July, 1953.) The field components and propagation characteristics for the TM mode are derived. The hollow waveguide is treated as a special case. Part 1: 328 of February.

621.372.8 1671
Propagation of Microwaves through a Cylindrical Metallic Guide Filled Coaxially with Two Different Dielectrics: Part 3—S. K. Chatterjee. (*Jour. Indian Inst. Sci.*, Section B, vol. 35, pp. 149–169; Oct., 1953.) The propagation characteristics for the TE_{01} and TM_{11} modes are derived. The results indicate that the phase velocity for a given mode can be adjusted to a pre-assigned value by a suitable choice of the dielectric constants and radii of the two media. In the case of the TE_{01} mode most of the energy is located in the medium having the higher dielectric constant. Part 2: 1670 above.

621.372.8 1672
Annular Resonant Slots in Dielectric-Filled Circular Waveguide—M. Cohn. (*Trans. I.R.E.* vol. MTT-1, pp. 39–44; Nov., 1953.) The variation of susceptance with frequency over the range 8.2–9.6 kmc was investigated experimentally for irises with annular slots of various dimensions, in polystyrene-filled waveguides operating in the TE_{11} mode. Relations between the resonance frequency and Q value and the slot dimensions are derived. Results are shown in graphs.

621.372.8 1673
Graphical Analysis of Measurements on Multiport Waveguide Junctions—S. Stein. (Proc. I.R.E., vol. 42, p. 599; March, 1954.) Extension of previous work [3191 of 1953 (Storer et al.)].

621.372.8 1674
On the Propagation Constant in Gentle Circular Bends in Rectangular Wave Guides—Matrix Theory—A. T. de Hoop. (*Jour. Appl. Phys.*, vol. 25, p. 136; Jan., 1954.) Correction to paper noted in 968 of April.

621.396.67.012.12:517.864 1675
Solution of the Integral Equation of a Linear Aerial—L. D. Bakhrakh. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 92, pp. 755–758; Oct. 1, 1953. In Russian.) The radiation-pattern function is related to the antenna current by a Fredholm-type equation of the first kind. The conditions are derived for the solutions to be in the form of (*a*) a finite sum of Mathieu harmonics and (*b*) a convergent series of Mathieu functions. The solution can be extended to the case of a plane antenna.

621.396.67.029.62:621.397 1676
Band III Television Aerials—F. R. W. Strafford. (*Wireless World*, vol. 60, pp. 181–184; April, 1954.) Requirements of antennas for receiving the proposed British transmissions in band III are discussed on the basis of the known performance of band-I antennas.

621.396.673.029.51 1677
Design and Adjustment of the Aerial at the National Long-Wave Station [at Allouis]—R. Chaste. (*Ann. Radioélect.*, vol. 8, pp. 301–312; Oct., 1953.) Design details are given of the folded λ/4 vertical antenna operating on 164 kc with bandwidth 20 kc [886 of March (Gaillard)]. Preliminary measurements were made on a model, scale 1:170.

621.396.674.3 1678
The Dipole and its Radiation—A. Spork. (*Elektrotech. u. Maschinenb.*, vol. 70, pp. 545–554; Dec. 15, 1953.) Theory introductory to a series of papers on antennas is presented.

621.396.674.33 1679
The Symmetrical Biconical Aerial with Arbitrary Cone Angle—L. Robin and A. Pereira-Gomes. (*Ann. Télécommun.*, vol. 8, pp. 382–390; Dec., 1953.) Schelkunoff's method (1049 of 1942) is extended to the general case, with the semi-angle of the cone, ψ, having any value in the range $0<\psi<\pi/2$. Numerical computations are made for $\psi=15$ degrees, 30 degrees, 45 degrees, 60 degrees, and 75 degrees. The corresponding normalized impedances at the end and at the center of the dipole are then calculated, taking into account the first two complementary internal waves and the first three external waves, in addition to the principal waves. Results are shown graphically.

621.396.676.012.12 1680
Electrically Small Antennas and the Low-Frequency Aircraft Antenna Problem—J. T. Bolljahn and R. F. Reese. (*Trans. I.R.E.*, vol. AP-1, pp. 46–54; Oct., 1953.) Quasi-static measurement techniques suitable for the determination of the radiation pattern and receiving sensitivity of electrically small antennas of both the electric- and magnetic-dipole types are described. These involve measurements on models in an electrostatic cage or an electrolyte tank. The theory, applications and limitations of these methods are discussed.

621.396.677 1681
A Simplified Calculation for Dolph-Tchebycheff Arrays—G. J. van der Maas. (*Jour. Appl. Phys.*, vol. 25, pp. 121–124; Jan., 1954.) A fuller account of work previously reported (634 of March).

621.396.677.3 1682
A Four-Terminal Network Theory for the Multi-Elements-Beam-Array with Reflector. (On an Adjustment Method of the Reflector and the Tailwire)—K. Nagai, R. Sato, and G. Sato. (*Tech. Rep. Tohoku Univ.*, vol. 17, pp. 131–147; 1953.) The antenna array is treated as a four-terminal network with input terminals at the feed points of the energized element and output terminals at the junction of the reflector and stub (tailwire), the stub being considered as a load. From the equations obtained, the array characteristics and the optimum stub length are calculated. Fair agreement with experimental values is obtained.

621.396.677.3.012.12 1683
Radiation Pattern Synthesis—E. A. Laport. (*RCA Rev.*, vol. 14, pp. 533–545; Dec., 1953.) A computational method is described for the synthesis of radiating arrays having zero radiation in specified directions. An extension of this method includes the solutions for radiation patterns from uniform current sheets of any specified aperture. Examples of radiation pattern shaping by using various primary arrays together with various secondary arrays are given.

621.396.677.7 1684
Passive Radiating Systems in Waveguides—M. L. Levin. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 91, pp. 807–810; Aug. 1, 1953. In Russian.) A generalization of results obtained earlier for the characteristics of radiation from a resonant slot in a waveguide. The generalization includes the case of an arbitrary nonresonant radiating system consisting of passive metal antennas and slots in waveguide walls.

621.396.677.85 1685
Propagation in an Artificial Dielectric—R. Fortet. (*Ann. Télécommun.*, vol. 8, pp. 361–366; Nov., 1953.) Periodic structures of the type used for microwave lenses are considered, comprising "free zones," which are dielectrically homogeneous, alternating with regions in which the dielectric material is partitioned by metal strips into "restricted zones." The simple case studied by Brillouin (1584 of 1949) is examined and a more generalized theory is developed.

AUTOMATIC COMPUTERS

681.142:[551.510.3:535.325 1686
An Analogue Computer for the Solution of the Radio Refractive-Index Equation—Johnson. (*See* 1760.)

681.142:621.385.832 1687
Automatic Beam Current Stabilization for Williams Tube Memories—R. J. Klein. (*Trans. I.R.E.*, vol. EC-2, pp. 8–11; Dec., 1953.) One of the storage points in the memory tube is used as a test point. At regular intervals this is cleared and recharged, and the output is sampled, the beam current being adjusted to keep the sampled output at a constant level. The whole operation takes 40 μs to complete. Circuit details are given.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.21 1688
A Note on calculating the Total Impedance of Impedors connected in Parallel—W. Boesch. (*Frequenz*, vol. 7, pp. 331–335; Nov., 1953.) Three methods, one involving use of a desk calculating machine, and another use of a linear nomogram, are compared and contrasted from the point of view of speed and accuracy.

621.3.018.783:621.376.3 1689
Nonlinear Distortion of a Random Signal. Application to Frequency Modulation—P. J. M. Clavier. (*Câbles & Transm.*, vol. 7, pp. 293–300; Oct., 1953.) Analysis is based on the instantaneous statistical properties of a random signal, the energy spectrum for a given bandwidth being determined from the appropriate correlation functions. Application of the method is illustrated in the case of FM when propagation time is considered. Since the calculation takes account of intermodulation products over a complete frequency band the method has advantages over that based on intermodulation of discrete frequencies.

621.314.22.018.75:621.318.1:538.221 1690
Pulse Permeability and Losses in Magnetic Materials subjected to Rectangular D.C. Pulses—Kinsele. (*See* 1826.)

621.318.435.3:621.375.3 1691
Transductor with High-Impedance Control-Winding Termination—W. Schmidt. (*Arch. elekt. Übertragung*, vol. 7, pp. 574–578; Dec., 1953.) The dynamic behavior of this transductor is represented by equations similar to those for a transductor with a low-impedance control-winding termination (2579 of 1953).

621.372:512.831 1692
Matrix Analysis of Linear Time-Varying Circuits—L. A. Pipes. (*Trans. I.R.E.*, PGCT-2, pp. 91–105; Dec., 1953.)

621.372+621.3.018.78](083.74) 1693
I.R.E. Standards on Circuits: Definitions of Terms in the Field of Linear Varying Parameter and Nonlinear Circuits, 1953—(Proc. I.R.E., vol. 42, pp. 554–555; March, 1954.) Standard 1953 IRE 4S1.

621.372.029.63/.64:621.315.212 1694
A Coaxial Magic-*T*—T. Morita and L. S. Sheingold. (*Trans. I.R.E.*, vol. MTT-1, pp. 17–23; Nov., 1953.) The basic design consists of a coaxial-line T junction including a shielded loop whose axis is located at the center of the T; the operation is similar to that of the hybrid

coil or the waveguide magic T. Two modifications are described, one using direct coupling between the loop and the loads. Standing-wave characteristics of experimental models for operation at wavelengths of about 10 cm are given; a swr<2 can be obtained over a frequency band of about 10 per cent. Application to phase measurements is described briefly.

621.372.221:621.385.029.63/.64 1695
Filter-Helix Traveling-Wave Tube: Part 1—The Filter Helix, a New Circuit Element for Traveling-Wave Amplifiers and Oscillators—Dodds and Peter. (*See* 1963.)

621.372.412 1696
The Properties and Manufacture of Piezoelectric Quartz Crystals—H. L. Downing. (*Jour. Brit. I.R.E.*, vol. 14, pp. 130–138; March, 1954.) Reprint. See 667 of March.

621.372.412:534.133 1697
Thickness-Shear and Flexural Vibrations of Contoured Crystal Plates—R. D. Mindlin and M. Forray. (*Jour. Appl. Phys.*, vol. 25, pp. 12–20; Jan., 1954.) Equations of motion expressing approximately and separately the thickness-shear and flexural modes are derived from data on the coupled modes for a plate of uniform thickness [1861 of 1951 and 2156 of 1952 (Mindlin)]. These equations are applied to the case of (*a*) a plate of double-wedge shape, and (*b*) a plate with bevelled edges. Results support the finding that thickness-shear motion is localized at the center, flexural motion at the edges.

621.372.412:534.133 1698
Thickness-Shear Vibrations of Piezoelectric Crystal Plates with Incomplete Electrodes—R. D. Mindlin and H. Deresiewicz. (*Jour. Appl. Phys.*, vol. 25, pp. 21–24; Jan., 1954.) The method developed in 1697 above (Mindlin and Forray) is applied to determine the influence of electrode length on thickness-shear frequencies. For a partially coated crystal these frequencies depend on the coupling between the coated and uncoated portions considered as separate systems.

621.372.412:534.133 1699
Suppression of Overtones of Thickness-Shear and Flexural Vibrations of Crystal Plates—R. D. Mindlin and H. Deresiewicz. (*Jour. Appl. Phys.*, vol. 25, pp. 25–27; Jan., 1954.) Analysis showing how electrodes coated on the surfaces of thickness-shear plates may be shaped so that only one resonance mode is excited.

621.372.413 1700
Derivation of Equivalence Theorem of a Weakly Coupled Electromagnetic Cavity-Resonator System—E. Ledinegg and P. Urban. (*Arch. elekt. Übertragung*, vol. 7, pp. 561–568; Dec., 1953.) A general proof of the equivalence between a system of weakly coupled cavities and a system of coupled resonant LCR circuits is derived from Maxwell's equations by means of a first-order perturbation calculation. The frequency dependence of the system is deduced from the equation of state. The case of two coupled cavity resonators is considered in an example.

621.372.413 1701
Calculation of the Resonant Properties of Electrical Cavities—S. Bertram. (Proc. I.R.E., vol. 42, pp. 579–585; March, 1954.) The method described is based on measurements made on a model in an electrolyte tank. Calculated values of resonance wavelength and Q are in good agreement with directly measured values.

621.372.414:517.5 1702
Characteristic Functions of Real Resonators—V. M. Vakhnin. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 91, pp. 779–782; Aug. 1, 1953. In Russian.) Orthogonal characteristic functions for a piecewise-inhomogeneous, one-dimensional system are considered in relation to resonators with energy losses at the boundaries. The analysis provides an explanation of previously published experimental results.

621.372.5 1703
An Iterative Method for Network Synthesis—R. E. Scott and R. L. Blanchard. (*Trans. I.R.E.*, PGCT-2, pp. 19–29; Dec., 1953.) The numerical method described is based on the complex critical frequencies of the insertion-loss function. Applications to networks with up to four poles and four zeros and to a crystal filter are illustrated.

621.372.5 1704
Synthesis of Transfer Functions with Poles Restricted to the Negative Real Axis—A. D. Fialkow and I. Gerst: L. Weinberg. (*Jour. Appl. Phys.*, vol. 24, pp. 1525–1527; Dec., 1953.) Comment on 1929 of 1953 and author's reply.

621.372.5:621.396 1705
Radio-Frequency Phase-Difference Networks: New Approach to Polyphase Selectivity—Cifuentes and Villard. (*See* 1910.)

621.372.5.018.78 1706
Quasi-Distortionless Filter Functions—J. L. Stewart. (*Trans. I.R.E.*, PGCT-2, pp. 39–54; Dec., 1953.) The quasi-distortionless filter network function is specified by expressing the input and output functions as time power series and equating their coefficients. Such filter functions are derived and discussed.

621.372.54 1707
The Direct Method of Filter and Delay Line Synthesis—M. J. E. Golay. (Proc. I.R.E., vol. 42, pp. 585–588; March, 1954.) Extension of previous work (1497 of 1946). "The direct method consists in postulating an indefinitely recurring network in which each recurring element is coupled to all elements of the network, and in determining the values of these elements and of their indefinitely extended couplings which will yield the exact phase characteristics desired within the exact pass band desired." Examples are given of the application of the method to the design of ideal band-pass filters.

621.372.54 1708
The Synthesis of Reactance Networks consisting only of Rejector Circuits in Series, by the Method of Partial Networks—T. O'Callaghan. (*Frequenz*, vol. 7, pp. 299–306; Oct., 1953.) The modification required in a filter network with given resonance and antiresonance frequencies in order to introduce a further prescribed resonance or antiresonance frequency is determined. Numerical examples illustrate the method, which is also applied to the reduction of the number of resonance frequencies and to a mechanical oscillator.

621.372.54 1709
The Historical Development of Filter Theories—W. Klein. (*Frequenz*, vol. 7, pp. 326–331; Nov., 1953.)

621.372.54 1710
Elementary Ladder-Type Filters—J. Oswald. (*Cables & Transm.*, vol. 7, pp. 325–358; Oct., 1953.) A system of classification of conventional ladder-type filters based on image parameters is described. An elementary filter is defined as a network with two prescribed image impedances and minimum attenuation. These are generally composed of conventional half-sections but in certain cases, particularly for band-pass filters, synthesis involves sections with three branched impedances which cannot be resolved into matched half-sections. Formulas with examples of networks for low-pass, high-pass, band-pass, and band-stop filters are given in an appendix.

621.372.54:621.314.7 1711
***RC* Active Filters**—J. G. Linvill. (Proc. I.R.E., vol. 42, pp. 555–564; March, 1954.) A type of filter is described in which the active element is a negative-impedance converter using transistors (2582 of 1953). The number of capacitors required is only equal to the total number of reactance elements in the corresponding LC filter. Theory is given, and tests on low-pass, high-pass, and band-pass filters are reported.

621.372.543.2 1712
Impedance Transformations in Band-Pass Filters—T. J. O'Donnell and E. M. Williams. (*Elect. Eng.*, vol. 72, p. 1105; Dec., 1953.) Digest of paper to be published in *Trans. AIEE*, vol. 72, 1953. The design of band-pass impedance-transforming filters is derived from the equivalence between an asymmetrical band-pass filter and a symmetrical section plus an ideal transformer.

621.372.543.2:621.375.221 1713
Design of a Simple Band-Pass Amplifier with Approximate Ideal Frequency Characteristics—W. E. Bradley. (*Trans. I.R.E.*, PGCT-2, pp. 30–38; Dec., 1953.) The application of the "pole and zero" method of synthesis is illustrated.

621.372.543.2:621.376.3 1714
F.M. Transient Response of Band-Pass Circuits—R. E. McCoy. (Proc. I.R.E., vol. 42, pp. 574–579; March, 1954.) A general formula is derived for the instantaneous deviation of output frequency resulting from a stop change of input frequency. Calculations are presented for large frequency shifts in a single-tuned circuit and for small frequency shifts in amplifiers with various numbers of tuned circuits. For small input steps near the middle of the pass band the transient response is the same for FM as for AM. Stagger tuning gives rise to overshoot, but the extent of this is small if the amount of stagger is no more than enough to provide maximally flat response in the steady state.

621.372.543.29 1715
One Design Method of Band-Pass Filter for V.H.F.—K. Nagai, R. Sato, and N. Saito. (*Technol. Rep. Tohoku Univ.*, vol. 17, pp. 120–130; 1953.) Design formulas for band-pass filters constructed of two-wire and coaxial cables are derived and numerical examples are given. The theoretical and experimental results are in fair agreement.

621.373:513.83 1716
"Hereditary" Dynamic Systems with Discontinuities ["à déferlement"]—T. Vogel. (*Ann. Télécommun.*, vol. 8, pp. 354–360; Nov., 1953.) The characteristics distinguishing nonlinear from linear oscillations are examined. A class of dynamic systems termed "discontinuous" ("à déferlement") is defined; cases are discussed for which Poincaré's topological methods can be used to determine the periodic solutions and the stability. Such systems are included in the larger class of "hereditary" systems defined by Volterra (i.e. systems whose operation is affected by their previous history). See also 630 of 1952.

621.373.4 1717
Single-Stage Phase-Shift Oscillator—W. Bacon. (*Wireless Eng.*, vol. 31, pp. 100–104; April, 1954.) A method of design is developed in which the phase-shifting network contains an arbitrary number of RC or CR cells and the value of the first series impedance is half that of the others. Simple formulas are derived for the network impedance and oscillation frequency.

621.373.5 1718
Transistor Oscillators—H. E. Hollmann. (*Arch. elekt. Übertragung*, vol. 7, pp. 585–591; Dec., 1953.) The self-oscillation conditions and frequency-deviation characteristics of a three-point transistor-oscillator are derived by analogy with the retarding-field oscillator, and a comparison is made with measured characteristics of point-contact and junction-type tran-

sistor oscillators. Various circuits are examined from the point of view of suitability of application of the two types of transistor. The discussion is limited to the frequency range below α cut off.

621.375.132.012 1719
The Audio-Frequency Negative-Feedback Amplifier as a Filter Circuit—F. Steiner and I. Riess. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 7, pp. 141–147; Nov./Dec., 1953.) A single-stage *RC* feedback amplifier is represented by an equivalent damped parallel *I C* circuit and a locus diagram for amplification is constructed. The method is applied to a two-stage amplifier, an expression analogous to form factor in filter theory being derived in terms of feedback ratio. This is useful for determining the effect of feedback on the frequency characteristic in af amplifier design.

621.375.2 1720
Bandpass Amplifiers—Z. Jelonek and R. S. Sidorowicz. (*Wireless Eng.*, vol. 31, pp. 84–99; April, 1954.) A design method is presented based on the required over-all gain, half-power bandwidth and center frequency of the pass band. The amplifier considered is composed of a number of identical groups each comprising one or more amplifying stages and having a maximum-flatness frequency response. Conditions for maximum possible gain are found. Regeneration due to anode-grid capacitances is discussed; the amount of distortion introduced into the frequency response is related to a regeneration coefficient α. When α is sufficiently large, oscillations occur. Stability conditions for single-tuned and stagger-tuned amplifiers are derived.

621.375.221 1721
Distributed Amplifier Theory—D. V. Payne. (PROC. I.R.E., vol. 42, pp. 596–598; March, 1954.) Discussion on 2612 of 1953.

621.375.222.029.3 1722
A New Transformerless Amplifier Circuit—K. Onder. (*Jour. Audio Eng. Soc.*, vol. 1, pp. 282–286; Oct., 1953.) Four tubes in a bridge circuit are driven in appropriate phase and amplitude so as to deliver equal power to a common load. No dc flows through the load; negative feedback is easily applied. A 9-w and an 18-w amplifier are described; response characteristics are flat over the audio range at all levels of operation. The 18-w unit weighs 4 pounds 3 ounces, including power supply and pre-amplifier.

621.375.3 1723
Magnetic Amplifier performs Analytical Operations—L. A. Finzi and R. A. Mathias. (*Elect. Eng.*, vol. 72, p. 1097; Dec., 1953.) Digest of paper to be published in *Trans. AIEE*, vol. 72, 1953.

621.395.665 1724
Audio Automatic Volume Control Systems—F. W. Roberts and R. C. Curtis. (*Jour. Audio Eng. Soc.*, vol. 1, pp. 310–316; Oct., 1953.) Applications of avc are noted and various circuits are reviewed. Methods of eliminating audible "thump" occurring on application of a direct control voltage, of deriving the control voltage, and of reducing time delay are discussed. Measurement techniques are outlined.

621.395.665.1 1725
The New Compressor-Expander System of the French Post Office—M. Lagarde, R. Blondé, and R. Derosier. (*Câbles & Transm.*, vol. 7, pp. 301–308; Oct., 1953.) The new unit has improved operating characteristics and is less than one quarter the size of the former type [306 of 1951 (Lagarde et al.)]. Two Si diodes replace the Cu_2O rectifiers in the potentiometer system, and two Ge diodes the double-diode tube in the detector unit. Other tubes are replaced by miniature types. Performance tests are reported.

GENERAL PHYSICS

530.112 1726
Proposal for a New Aether Drift Experiment—H. L. Furth. (*Nature* (*London*), vol. 173, pp. 80–81; Jan. 9, 1954.) An experiment using standing microwaves is outlined.

535.1:537.228 1727
Deductions regarding the Constitution of Light Energy—J. Stark. (*Z. Phys.*, vol. 136, pp. 221–223; Nov. 17, 1953.) A discussion based on the results of experiments previously reported (1292 of 1953). The field associated with a particle of light, as defined by Planck, is nonoscillatory, the frequency associated with the light being due to the circulation of em energy in the particle.

535.37:537.228 1728
The Mechanism of Electroluminescence: Part 2—Applications to the Experimental Facts—D. Curie. (*Jour. Phys. Radium*, vol. 14, pp. 672–686; Dec., 1953.) Theory previously developed (1373 of May) is used to interpret observed phenomena, including temperature and frequency effects, fluctuations of brightness corresponding to field alternations, initial rise of brightness, and surface effects. Good crystallization, abundance of luminescence centers, and donor levels, and absence of other defects are conditions favoring the production of electroluminescence. Various theories of the phenomenon are compared in an appendix.

535.41+621.372.2]:621.3.012 1729
On a Property of a Family of Equiangular Spirals and its Application to Some Problems of Wave Propagation—Landecker. (*See* 1667.)

535.42:538.566 1730
The Diffraction of Waves in passing through an Irregular Refracting Medium—J. A. Fejer. (*Proc. Roy. Soc. A*, vol. 220, pp. 455–471; Dec. 22, 1953.) "A relation is derived between the angular power spectrum of waves emerging from a thin diffracting screen random in two dimensions and the auto-correlation function describing the irregularities of the field as it emerges from the diffracting screen. The special case of an isotopic screen characterized by an auto-correlation function which depends only on distance and not on direction is discussed for normal and oblique incidence. Multiple scattering of waves caused by volume irregularities of the dielectric constant is considered. The angular power spectrum and the auto-correlation function describing the irregularities of the diffraction field caused by multiple scattering in a thick slab are determined. The results are compared with those obtained by Hewish [1284 of 1952] for a thin phase-changing screen. The results are discussed in terms of problems arising in the study of radio-wave propagation."

537.213 1731
A Note on Uniqueness Proofs for Boundary-Value Problems in Potential Theory and Steady Heat Conduction—M. E. Rayner. (*Quart. Jour. Mech. Appl. Math.*, vol. 6, part 4, pp. 385–390; Dec., 1953.) By allowing the boundary to tend to infinity in two directions, and considering the problem in a semi-infinite region, it is possible to prove uniqueness by classical methods for a very large class of boundary conditions.

537.311.5:517.9 1732
Variational Methods for Problems in Resistance—J. F. Carlson and T. J. Hendrickson. (*Jour. Appl. Phys.*, vol. 24, pp. 1462–1465; Dec., 1953.) Schwinger's variational methods are used to calculate upper and lower limits for the resistance of a cylinder with a specified transverse distribution of potential at one end.

537.311.62:517.392 1733
The Evaluation of the Surface Impedance in the Theory of the Anomalous Skin Effect in Metals—A. N. Gordon and E. H. Sondheimer. (*Appl. Sci. Res.*, vol. B3, pp. 297–304; 1953.) A method is described for obtaining series expansions for the integrals which represent the surface impedance.

537.52 1734
Note on the Quantitative Development of the [gas] Discharge Characteristic—E. Pfender. (*Z. angew. Phys.*, vol. 5, pp. 450–453; Dec., 1953.)

537.52:551.594.223 1735
Investigation of Ball Lightning by means of Models—H. Nauer. (*Z. angew. Phys.*, vol. 5, pp. 441–450; Dec., 1953.) Experiments with gas discharges are described.

538.2 1736
U.S.S.R. Research on Magnetism—S. Rosenblum. (*Nuovo Cim.*, Supplement to vol. 10, pp. 441–458; 1953. In French.) A brief historical survey with comprehensive bibliography.

538.214 1737
Effect of the Surface on the Magnetic Properties of an Electron Gas—F. S. Ham. (*Phys. Rev.*, vol. 92, pp. 1113–1119; Dec., 1, 1953.) The energy levels of free electrons confined in a finite cylindrical box with a uniform axial magnetic field are computed using the WKB approximation; the form of approximation appropriate to various boundary conditions is discussed. The results are used to calculate the susceptibility of the system.

538.3 1738
On Reciprocity Theorems in Electromagnetic Theory—T. H. Crowley. (*Jour. Appl. Phys.*, vol. 25, pp. 119–120; Jan., 1954.) Five different types of sources of em fields are considered, and the corresponding reciprocity theorems stated.

538.3 1739
A Scalar Representation of Electromagnetic Fields—H. S. Green and E. Wolf. (*Proc. Phys. Soc.*, vol. 66, pp. 1129–1137; Dec. 1, 1953.) It is shown that in a region which is free from currents and charges any em field may be rigorously derived from a single, generally complex, scalar wave function. The momentum density and energy density defined in terms of this function differ from those given by the usual expressions, but the differences disappear on integration over any arbitrary macroscopic domain.

538.5 1740
Induction of Currents by Moving Charges—Ya. N. Fel'd. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 93, pp. 447–450; Nov. 21, 1953. In Russian.) An analysis is made of the field due to a charge moving between a pair of parallel disk electrodes, connected externally by a conductor. A general expression for the current flowing through the conductor is derived in terms of the electrode radius, r_0, and separation, a, and the velocity of the charge, u. A periodic discharge could be obtained with particular values of r_0, a, and u.

538.561:662.2 1741
Electromagnetic Waves emitted on Detonation of Explosives—H. Kolsky. (*Nature* (*London*), vol. 173, p. 77; Jan. 9, 1954.) Small electrical disturbances, detected incidentally while using explosives to produce sharp stress pulses, were later investigated separately. The effect was displayed on a high-speed cro. The potential of a wire probe reached a maximum after about 50 μs and then decayed to zero. The effect is probably due to the production of positive and negative ions having different mobilities.

538.566 1742
Vector Modifications inherent in the Doppler Effect for Waves Propagated in a Dielectric Medium—M. Risco. (*Jour. Phys. Radium*, vol. 14, pp. 657–662; Dec., 1953.) The propagation of plane em waves in a moving medium is

characterized by four vectors, as compared with two for a stationary medium—because the directions of the electric and magnetic fields no longer coincide with the respective inductions. Relations between these vectors are derived and their physical significance is discussed.

538.566 1743
The Electromagnetic Transmission Characteristics of the Two-Dimensional Lattice Medium—H. S. Bennett. (*Jour. Appl. Phys.*, vol. 24, pp. 785–810; June, 1953.) A theoretical investigation is made of systems with obstacles arranged to form a regular lattice in planes perpendicular to the direction of propagation, the obstacles having elliptical cross-section. An examination is made of the influence of obstacle size, shape, and spacing and of the wave polarization on the transmission characteristics. No restriction is introduced as regards the size of the obstacle in relation to λ. The theory is relevant to problems of microwave lenses and transmission circuits of traveling-wave tubes.

538.566:535.42 1744
Microwave Diffraction Measurements in a Parallel-Plate Region—R. V. Row. (*Jour. Appl. Phys.*, vol. 24, pp. 1448–1452; Dec., 1953.) Diffraction of a plane or a cylindrical wave by wedges, covered with 0.003-inch silver foil, was investigated experimentally by a probe method. The wedge, the 3.185-cm-λ line source and the probe were located between two large parallel duralumin sheets spaced $\frac{1}{2}$-inch apart. The measured values of field strength are compared with values derived from previously published theory.

538.566:538.311 1745
Radiation from a Line Source Adjacent to a Conducting Half Plane—J. R. Wait. (*Jour. Appl. Phys.*, vol. 24, pp. 1528–1529; Dec., 1953.) Expressions for the field are obtained in terms of tabulated functions by using an alternative method of analysis to that of Harrington (2967 of 1953).

538.566.2 1746
The Fields of a Line Source of Current over a Stratified Conductor—J. R. Wait. (*Appl. Sci. Res.*, vol. B3, pp. 279–292; 1953.) Expressions for the electric and the magnetic fields are derived, assuming the wire to be infinitely long and parallel to the conductor. At large distances the results can be generalized to apply to any source of horizontally polarized waves situated on the surface of a half-space consisting of an arbitrary number of layers.

538.569.4.029.64 1747
Absorption of Microwaves in Gases—Krishnaji and P. Swarup. (*Jour. Appl. Phys.*, vol. 24, p. 1525; Dec., 1953.) Absorption coefficients in the 3-cm wave band were determined for ammonia, ethyl alcohol and acetaldehyde at a pressure of 745 mm Hg, and for methyl iodide and acetone at 20 cm. Hg and 25 cm Hg respectively.

539.23:537.533.8/.9 1748
The Decomposition of Thin Films on Bombardment with Slow Electrons—D. A. Wright and J. Woods. (*Proc. Phys. Soc.*, vol. 66, pp. 1073–1086; Dec. 1, 1953.) The threshold energies of the bombarding electrons were determined for the onset of electron absorption, decomposition and secondary emission. The films investigated included BaO, $BaCl_2$, $BaSO_4$ and the alkali halides. The experiments were carried out using diode or triode tube systems, with an oxide-coated cathode operated at 400 degrees C. as the source of the bombarding electrons.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5:621.396.96 1749
Radio Astronomy: Part 3—Radiolocation of Meteors—W. Dieminger. (*Arch. elekt. Übertragung*, vol. 7, pp. 555–560; Dec., 1953.) A review of measurement techniques and results with a note on local ionization effects of meteors. Part 2: 1407 of May (Siedentopf).

523.5:621.396.96 1750
The Effect of Radar Wavelength on Meteor Echo Rate—V. R. Eshleman. (*Trans. I.R.E.*, vol. AP-1, pp. 37–42; Oct., 1953.) Theory is given which includes the effects of high electron density, the linear rate of trail formation, and the initial column radius on the meteor echo power and hence on the number of echoes received. The theories of Lovell and Clegg (2782 of 1948) and others are discussed critically. A comparison with experimental results is made.

523.7:550.38 1751
A Note on Solar-Terrestrial Relationships—N. C. Gerson. (*Jour. Atmos. Terr. Phys.*, vol. 4, pp. 341–342; Jan., 1954.) Corpuscular emission from the invisible solar disk and deflection of the particles into certain trajectories could account for the lack of correlation between particular solar and geophysical phenomena.

523.746 1752
Identification of Sunspots—H. Alfvén. (*Tellus*, vol. 5, pp. 423–445; Nov., 1953.) Previous work (3104 of 1948) indicates a statistical correlation between spots on opposite hemispheres during successive sunspot cycles. Analysis of sunspots between latitude 4 degrees S and 4 degrees N appears to support this view. A number of sunspots are identified as being produced by the same original disturbance in the solar core; the mechanism is explained on the basis of the magneto-hydrodynamic theory of sunspots.

523.854:621.396.822 1753
Position and Identification of a Bright Extended Radio Source in Gemini—J. E. Baldwin and D. W. Dewhirst. (*Nature (London)*, vol. 173, pp. 164–165; Jan. 23, 1954.) More accurate observations are reported on a previously detected source [121 of 1952 (Ryle et al.)].

523.754:621.396.822 1754
The Radio Brightness Distributions over Four Discrete Sources of Cosmic Noise—B. Y. Mills. (*Aust. Jour. Phys.*, vol. 6, pp. 452–470; Dec., 1953.) The experimentally determined isophotes, obtained from measurements with a variable-aerial-spacing radio interferometer at 101 mc, are compared with the optical features of the nebulae.

523.854:621.396.822 1755
Galactic Radiation at Radio Frequencies: Part 5—The Sea Interferometer—J. G. Bolton. and O. B. Slee. (*Aust. Jour. Phys.*, vol. 6, pp. 420–433; Dec., 1953.) The factors governing the interference pattern of a Lloyd's-mirror type of interferometer at frequencies between 40 and 400 mc are discussed and three systems for reducing the effect of background noise are described. Part 4: 1905 of 1952 (Bolton and Westfold).

523.854:621.396.822:550.510.535 1756
Galactic Radiation at Radio Frequencies: Part 6—Low-Altitude Scintillations of the Discrete Sources—J. G. Bolton, O. B. Slee, and G. J. Stanley. (*Aust. Jour. Phys.*, vol. 6, pp. 434–451; Dec., 1953.) Scintillations of four discrete sources at altitudes from 0 degrees to 10 degrees were observed at frequencies in the range 40 to 300 mc. A strong correlation is established between the occurrence of scintillations and sporadic E. Part 5: 1755 above (Bolton and Slee).

550.385:551.55:551.510.535 1757
Note on Geomagnetic Disturbance as an Atmospheric Phenomenon—E. H. Vestine. (*Jour. Geophys. Res.*, vol. 58, pp. 539–541; Dec., 1953.) Possible relations between ionospheric winds and atmospheric electric currents associated with magnetic storms are briefly discussed.

550.385:551.556 1758
The Immediate Source of the Field of Magnetic Storms—E. H. Vestine. (*Jour. Geophys. Res.*, vol. 58, pp. 560–562; Dec., 1953.) Results of a comparison of mean values of initial phase of magnetic storms at Huancayo, Cheltenham (Md.), Honolulu and San Juan (Porto Rico), indicate that major immediate sources of magnetic-field changes during storms in all phases are within or near the atmospheric region. *See* also 1757 above.

551.510.3:535.325 1759
A Statistical Survey of Atmospheric Index-of-Refraction Variation—C. M. Crain, A. W. Straiton, and C. E. von Rosenberg. (*Trans. I.R.E.*, vol. AP-1, pp. 43–46; Oct., 1953.) Survey of microwave measurements made at heights of 2000–25,000 feet over the Pacific Ocean, California, and Ohio, using an airborne refractometer described by Crain and Deam (3493 of 1952).

551.510.3:535.325]:681.142 1760
An Analogue Computer for the Solution of the Radio Refractive-Index Equation—W. E. Johnson. (*Jour. Res. Nat. Bur. Stand.*, vol. 51, pp. 335–342; Dec., 1953.) A computer in use at the CRPL comprises basic computation circuits incorporated in a bridge circuit. A 10-turn potentiometer is modified so as to give an approximately exponential variation of the resistance with the shaft rotation, corresponding to the curve of the saturated-water-vapor term in the equation.

551.510.535 1761
Irregularities in the Ionosphere—R. Roy and J. K. D. Verma. (*Jour. Geophys. Res.*, vol. 58, pp. 473–485; Dec., 1953.) Electron clouds in both the E and the F regions have been detected at vertical incidence by means of the equipment described by Banerjee and Roy (1677 of 1953). Photographs of typical echo records are shown; the power and persistence of the echoes are discussed.

551.510.535 1762
The Problem of Locating Inhomogenities in the Ionosphere—A. A. Gorozhankina. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 93, pp. 459–461; Nov. 21, 1953. In Russian.) Vertical soundings were made using two frequencies, $f_1=4.3$ mc and $f_2=6.4$ mc, chosen so that f_1 was reflected at the F_1 layer or the lower part of the F_2 layer and f_2 near the middle of the F_2 layer. The correlation factors ρ between pairs of the four parameters defined by $k=(\overline{\Delta R}/\overline{\Delta t})/\overline{R}$, derived from the amplitudes R of the ordinary and the extraordinary f_1 and f_2 signals recorded over periods t of a few minutes, are calculated and their significance is discussed. A high degree of correlation ($\rho > 0.9$) between the four parameters indicates that the inhomogeneous region lies in the path common to the f_1 and f_2 signals.

551.510.535 1763
Recombination and Diffusion and Spread Echoes from the Ionosphere—T. L. Eckersley. (*Proc. Phys. Soc.*, vol. 66, pp. 1025–1038; Dec. 1, 1953.) The inclusion of a diffusion term in the differential equation of layer formation, to account for ionization due to the influx of charged particles, leads to a solution in terms of a doubly periodic elliptic function. This solution indicates the possibility of a periodic variation of ionic density within the F layer such as to account for spread echoes which have been observed. Layers produced by ultraviolet radiation are many wavelengths thick, those due to particle ionization are only two or three wavelengths thick and are embedded in the background of ionization. The existence of auroral curtains provides further evidence of ionization by particles.

551.510.535 1764
Origin of the E Layer of the Ionosphere—E. Bauer and Ta-You Wu. (*Phys. Rev.*, vol. 92, pp. 1101–1105; Dec. 1, 1953.) The electron

densities for the E and F_1 layers resulting from photoionization of molecular and atomic oxygen are calculated on the basis of the work by Moses and Wu on the distribution of O_2 and of O in and above the oxygen dissociation layer (129 of 1952 and 102 of 1953). While the results are in reasonable agreement with observations for the E layer, it is not possible to account for the observed value of recombination coefficient for the F_1 layer by any theory based on processes involving only oxygen atoms and molecules.

551.510.535 1765
Method of Determining the True Height of the Ionospheric Layers: Part 2—Application of the Exact Value of the Refractive Index (Ordinary-Ray Case)—É. Argence and M. Mayot. (*Jour. Geophys. Res.*, vol. 58, pp. 493–496; Dec. 1953. In French.) Values obtained using the exact value of the refractive index (Appleton-Hartree) are somewhat higher than those obtained with the approximate formulas used in part 1 (3290 of 1953).

551.510.535 1766
Recombination in the F Layers—A. L. Stewart. (*Nature*, (*London*), vol. 173, p. 165; Jan. 23, 1954.) The mechanism proposed by Banerji (1058 of April) is critically discussed.

551.510.535:551.594.12 1767
Further Discussion of Kelso's Paper on a Method for Determination of the Distribution of Electron Density in the Ionosphere—L. Kraus. (*Jour. Geophys. Res.*, vol. 58, pp. 551–553; Dec., 1953.) Comment on 1997 of 1953 (Manning).

551.510.535:551.594.5:550.384 1768
Correlation of Magnetic, Auroral, and Ionospheric Variations at Saskatoon—J. H. Meek. (*Jour. Geophys. Res.*, vol. 58, pp. 445–456; Dec., 1953.) Analysis of measurements made between December, 1951 and April, 1952. A connection exists between the maximum elevation of auroral light above the northern horizon and the maximum amplitude of variation of H. Some types of sporadic-E reflecting layers appear more frequently during geomagnetic disturbances.

551.594.223:537.52 1769
Investigation of Ball Lightning by means of Models—H. Nauer. (*Z. angew. Phys.*, vol. 5, pp. 441–450; Dec., 1953.) Experiments with gas discharges are described.

551.594.5:551.510.535:551.55 1770
On the Production of Glow Discharges in the Ionosphere by Winds—O. R. Wulf. (*Jour. Geophys. Res.*, vol. 58, pp. 531–538; Dec., 1953.) A source of excitation of the airglow and aurora may be found in the potential difference generated by zonal ionospheric winds cutting the earth's magnetic field. A return path of current flow or glow discharge in the higher ionosphere appears plausible. See also 3790 of 1945.

551.594.5+550.385]621.396.96 1771
Radio Echoes observed during Aurorae and Geomagnetic Storms using 35 and 74 Mc/s Waves simultaneously—L. Harang and B. Landmark. (*Jour. Atmos. Terr. Phys.*, vol. 4, pp. 322–338; Jan., 1954.) Observations made at Kjeller in 1952 and at Tromsoe during early 1953 are reported. Echoes are only received when the antenna is directed towards the North; the echo amplitude has a maximum value when the antenna is directed horizontally, and decreases rapidly with increased elevation of the antenna. The results indicate that there is no correlation between the position of visible aurorae and echo range but that there is a correlation between the occurrence of high E_s layer critical frequency and the reception of echoes. During geomagnetic storms echoes on 74 mc are observed only during the more more severe phases. Scattering from the ground or sea reflected back via an intense E_s layer is advanced in explanation of the observations.

LOCATION AND AIDS TO NAVIGATION

621.396.932+621.396.969.3 1772
Radio in the Service of the Seaman—C. V. Robinson. (*Proc. IEE*, part I, vol. 101, pp. 27–28; Jan., 1954.) Chairman's address, I.E.E. Southern Center. Communications and navigational aids are reviewed.

621.396.96:551.578 1773
Polarization of Radar Echoes from Meteorological Precipitation—I. M. Hunter. (*Nature* (*London*), vol. 173, pp. 165–166; Jan. 23, 1954.) A "circularly polarized" radar for operation on a wavelength of 3.2 cm has been constructed, in which the polarization parameters are accurately known and controlled (voltage ellipticity ratio 0.99 on transmission and 0.95 on reception) and spurious signals due to departure from circularity are eliminated by means of a grating in the waveguide which separates the incoming signals into two orthogonal components. Error signals can be reduced to a level of −50 db for point targets. Measurements on various types of precipitation are tabulated and discussed.

621.396.968:551.578.1 1774
Rain Clutter Measurements with C. W. Radar Systems Operating in the 8-mm Wavelength Band—D. G. Kiely. (*Proc. IEE*, part III, vol. 101, pp. 101–108; March, 1954.) Measurements were made using (*a*) a single antenna for transmission and reception, with polarization duplexing, and (*b*) separate antennas for transmission and reception, without duplexing. The amplitude of the clutter was given in terms of decibels below tube power level and was plotted against rainfall rate. The clutter was about 35 db greater in case (*a*) than in case (*b*). The maximum useful separation between the centers of the transmission and reception antennas is about 2 feet 6 inches. No definite result was obtained for reduction of clutter by use of the orthogonal polarization technique, but the reduction reported by other observers was generally confirmed.

621.396.969.13 1775
Height Indicator—P. H. Leidier. (*Ann. Radioélect.*, vol. 8, pp. 313–317; Oct., 1953.) The principle of operation of the height indicator Type IS 330 is described. It is associated with 10-cm radar equipment Type ER 221 measuring azimuth and range. The cro presentation shows range as abscissa and height relative to the station as ordinate. Sources of error are considered and the accuracy of the system is determined. A block diagram of the unit is given.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215:535.37 1776
Effect of Illumination Intensity, Field Strength and Temperature on the Dark Current and Photocurrent of Phosphors—H. Gobrecht, D. Hahn, and H. J. Kösel. (*Z. Phys.*, vol. 136, pp. 285–292; Dec. 8, 1953.) An experimental investigation of various phosphors is reported. The photocurrent increases linearly with the illumination intensity, at low intensities, and exponentially with electric field strength. It increases with temperature without exhibiting saturation in phosphors with a bimolecular luminescence mechanism, but reaches saturation early, or remains constant, in phosphors with a monomolecular mechanism. Results for both photocurrent and dark current are shown graphically.

535.215:546.682.231 1777
Photoconductivity of Indium Selenide—D. E. Bode and H. Levinstein. (*Jour. Opt. Soc. Amer.*, vol. 43, pp. 1209–1210; Dec., 1953.) Results are given of measurements of the spectral response of evaporated films at 20 degrees C. and −78 degrees C.

535.215:546.817.221 1778
Photoelectromagnetic and Photoconductive Effects in Lead Sulphide Single Crystals—T. S. Moss. (*Proc. Phys. Soc.*, vol. 66, pp. 993–1002; Dec. 1, 1953.) Bulk photoeffects have been observed. The quantum efficiency is approximately equal to unity between 0.9 and 2.9 μ and falls rapidly at 3 μ. The optical activation energy is 0.41 ev. Carrier lifetimes, which are roughly proportional to the square of the resistivity, lie between 6×10^{-10} and 9×10^{-6} sec.

535.215.1:537.311.33 1779
Sensitization of the Internal Photoeffect in Semiconductors by Chlorophyll and Similar Pigments—E. K. Putseiko and A. N. Terenin. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 90, pp. 1005–1008; June 21, 1953. In Russian.) The spectral response of ZNO is altered by adsorption of chlorophyll so that in addition to the 3600–3700 Å maximum, two additional maxima occur in the visible region, near 4300 Å and between 6600 and 6700 Å. Similar results were obtained with other pigments. Other semiconductors investigated include PbO, ZnS, CdS, CdI and PbI_2. The chlorophyll precipitates themselves did not show any photoeffects.

535.37 1780
Brightness Waves and Transitory Phenomena in the Quenching of Luminescence by Alternating Electric Fields—G. Destriau. (*Jour. Appl. Phys.*, vol. 25, pp. 67–71; Jan., 1954.) An experimental study was made of the variations of brightness of sulphide phosphors excited by means of a half-wave X-ray generator and subjected to a field alternating at the same frequency (50 cps) as that of the X-ray generator. Secondary brightness maxima were observed lagging a quarter of a cycle behind the normal X-ray-excited maxima. The amplitude of the secondary maxima varied rapidly from cycle to cycle immediately after application of the field.

535.37 1781
Variations in the Decay of Phosphorescence with Frequency of Applied Electric Field—K. W. Olson. (*Phys. Rev.*, vol. 92, p. 1323; Dec. 1, 1953.) Results of measurements on (Zn, Cd)S-Cu phosphors at 80, 800, and 5000 cps are reported.

535.37 1782
Mechanism of Impurity Poisoning in the Luminescence of Zinc Sulfide Phosphors with Manganese Activator—P. H. Bube, S. Larach, and R. E. Shrader. (*Phys. Rev.*, vol. 92, pp. 1135–1139; Dec. 1, 1953.) Poisoning effects of Fe, Co, and Ni were studied. The relative importance of three possible poisoning mechanisms considered depends on the activator and poison proportions and on the energy of the exciting radiation.

535.37 1783
Electroluminescence in Thin Films of Zns:Mn—R. E. Halsted and L. R. Koller. (*Phys. Rev.*, vol. 93, pp. 349–350; Jan. 15, 1954.) No luminescence is produced by application of a unidirectional field below breakdown strength, apart from flashes on application and removal of the field. With alternating fields, the brightness varies at a frequency twice that of the applied field. The dependence of the effect on temperature and frequency is investigated.

535.37 1784
The Penetration of Electrons into Luminescent Material—W. Ehrenberg and J. Franks. (*Proc. Phys. Soc.*, vol. 66, pp. 1057–1066; Dec. 1, 1953.) The penetration of electrons of energies between 10 and 40 kev into single crystals of Tl-activated iodides of K, Rb and Cs, and of tungstates of Ca and Cd, and a luminescent plastic, was measured on

microphotographs of the luminous region. The diameter of the electron beam was $<0.75\ \mu$.

535.37 1785
The Manganese Emission in ABF_3 Compounds—H. A. Klasens, P. Zalm, and F. O. Huysman. (*Philips Res. Rep.*, vol. 8, pp. 441–451; Dec., 1953.) Investigation of the fluorescence under cathode-ray excitation of perovskite-type compounds activated with Mn. The A metals used were Na, K, Rb, and Cs, the B metals Mg, Zn, Cd, Ca, and Sr. A tentative explanation is given of the shift from orange to green occurring with increase of the Mn-F distance.

535.37 1786
Electroluminescent Zinc Sulfide Phosphors—H. H. Homer, R. M. Rulon, and K. H. Butler. (*Jour. Electrochem. Soc.*, vol. 100, pp. 566–571; Dec., 1953.) The preparation of phosphors giving blue, green and yellow electroluminescence is described. The activation of the green and blue phosphors is produced by Cu together with Pb, that of the yellow phosphors by Cu with Pb and Mn. A controlled amount of chloride is included in all the phosphors.

535.37:537.311.33 1787
Electroluminescence of Insulated Particles—L. Burns. (*Jour. Electrochem. Soc.*, vol. 100, pp. 572–579; Dec., 1953.) An examination is made of possible mechanisms whereby luminescence could be produced in a region on or near the surface of a semiconductor particle embedded in a dielectric and subjected to an electric field. The localization of the field, the presence of electrons in the conduction band, the color shift with frequency in some phosphors, and the efficiency of the phosphor are discussed.

536.587:548.55:546.289 1788
Temperature Regulator for Germanium Metallurgy—G. Lehmann and C. Meuleau. (*Onde élect.*, vol. 33, pp. 678–683; Dec., 1953.) The importance of temperature control in the production of single crystals for Ge diodes and triodes is emphasized. In the system described the supply to an induction furnace is controlled by means of a thermometric resistance bridge and a servo motor. The temperature stability achieved is within $\frac{1}{8}$ degrees C. at 930 degrees C., for mains voltage variations up to 10 per cent.

537.226 1789
Origin of Ferroelectricity in Barium Titanate and other Perovskite-Type Crystals—H. D. Megaw. (*Acta Cryst.*, *Camb.*, vol. 5, part 6, pp. 739–749; Nov. 10, 1952.)

537.226.2:546.23 1790
The Dielectric Constant of Amorphous Selenium at Wavelengths of 1 cm and 3 cm—Y. Klinger and E. W. Saker. (*Proc. Phys. Soc.*, vol. 66, p. 1117; Dec. 1, 1953.) A cylindrical-cavity resonator operating in the H_{01} mode was used for measurements. The weighted mean value obtained is 6.37_6.

537.228.1:548.0 1791
Dynamic Determination of Elastic and Piezoelectric Coefficients—R. Bechmann. (*Telefunken Ztg.*, vol. 26, pp. 353–365; Dec., 1953.) The determination of the coefficients of piezoelectric crystals from measurements of resonance frequency, crystal dimensions, density and dynamic capacitance is discussed in detail; crystals of various structures are considered.

537.311.31 1792
The Propagation of Electrons in a Strained Metallic Lattice—S. C. Hunter and F. R. N. Nabarro. (*Proc. Roy. Soc. A.*, vol. 220, pp. 542–561; Dec. 22, 1953.) The problem is studied using a perturbation technique in which the perturbing potential is proportional to the elastic strain rather than to the displacement. For the approximation of nearly free electrons, the resulting potential depends only on the Fermi energy of the electrons and not on their interaction with the ionic lattice. In a higher approximation this interaction is taken into account. The method is used to estimate the resistivity produced by dislocations of edge and screw types in Na and Cu.

537.311.31 1793
Electrical Conductivity of Metals at High Current Density—E. S. Borovik. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 91, pp. 771–774; Aug. 1, 1953. In Russian.) Measurements were made using both steady currents at very low temperatures and current pulses of densities of the order of 10^6 A/cm² at 20.4 degrees and 78 degrees K. Results indicate that Pt, W, and Cu obey Ohm's law at densities up to $5–8\times10^6$ A/cm²; large deviations were found for Bi at $0.5–1\times10^6$ A/cm².

537.311.31 1794
The Extra-Resistivity due to Vacancies in Copper, Silver and Gold—P. Jongenburger. (*Appl. Sci. Res.*, vol. B3, pp. 237–248; 1953.)

537.311.33 1795
Some Problems in the Diffusion of Minority Carriers in a Semiconductor—S. Visvanathan and J. F. Battey. (*Jour. Appl. Phys.*, vol. 25, pp. 99–102; Jan., 1954.) "The exact solutions to the problems of the diffusion of minority carriers involved in the measurement of surface recombination velocity in a semiconductor with a sample geometry bounded by two infinite planes are presented. The reduction of the exact solutions to simple forms used in the analysis of experimental data is shown."

537.311.33 1796
Contribution to the Theory of Conduction in Semiconductors—W. H. Isay. (*Ann. Phys.*, *Lpz.*, vol. 13, pp. 327–348; Dec. 15, 1953.) Given the conductivity rise and decay curves of a CdS crystal, and assuming a three-energy-band model, the various parameters of the associated system of differential equations can be calculated.

537.311.33 1797
Effect of Traps on Carrier Injection in Semiconductors—H. Y. Fan. (*Phys. Rev.*, vol. 92, pp. 1424–1428: Dec. 15, 1953.) Two effects are considered, (*a*) that on photoconductivity, and (*b*) that on the spread of excess carrier concentration under an applied field. Trapped minority carriers, by causing an increase in majority carrier concentration, give rise to increased photoconductivity which may vary nonlinearly with light intensity and have a very long time constant. An expression is derived for the distribution of excess minority carriers as a function of time and space, neglecting diffusion. Results are in agreement with observations.

537.311.33 1798
New Semiconducting Compounds: Part 2—H. Welker. (*Z. Naturf*,. vol. 8a, pp. 248–251; April, 1953.) Measurements of conductivity as a function of temperature are reported for specimens of the crystalline compounds InSb, GaSb, and AlSb; the widths of the respective forbidden energy bands are determined. Characteristics are shown of the rectifier effect in AlSb, GaSb, GaAs, and InP and of the transistor effect *InP*. Part 1: 1454 of May.

537.311.33c539.165:621.311.6 1799
The Electron-Voltaic Effect in *p-n* Junctions induced by Beta-Particle Bombardment—F. Rappaport. (*Phys. Rev.*, vol. 93, pp. 246–247; Jan. 1, 1954.) Current/voltage characteristics are shown for Si-alloy wafer-type units bombarded from a 50-millicurie Sr^{90}-Y^{90} source. One of the units has been used as power supply for an af transistor oscillator. The efficiency of conversion of the radioactive power is about 0.4. Ge units were also investigated.

537.311.33:546.23 1800
Frequency Dependence of the Electrical Conductivity of Polycrystalline Selenium—H. Gobrecht and H. Hamisch. (*Z. Phys.*, vol. 136, pp. 234–247; Nov. 17, 1953.) The conductivity of Se layers deposited on glass by evaporation was measured at frequencies up to 14 mc and at temperatures between −25 degrees and +50 degrees C. The results are shown graphically and are discussed in relation to the Maxwell-Wagner theory for a capacitor with two different dielectrics.

537.311.33:546.23:548.55 1801
Electrical Properties of Selenium: Part 3—Microcrystalline Selenium Metal Doped—H. W. Henkels and J. Maczuk. (*Jour. Appl. Phys.*, vol. 25, pp. 1–11; Jan., 1954.) Measurements were made on two crystalline forms of pure Se and on samples with small amounts of various metals added. Thermoelectric power and the dc and 200-mc resistivities are studied as functions of temperature and a theoretical energy-band model for Se is developed. Part 2: 1335 of 1952 (Henkels).

537.311.33:[546.28+546.289 1802
Mobility of Impurity Ions in Germanium and Silicon—J. C. Severiens and C. S. Fuller. (*Phys. Rev.* vol. 92, pp. 1322–1323; Dec. 1, 1953.) The diffusion constants of Li in Si and Ge, and of Cu in Ge, as calculated from the experimentally determined mobilities, are in agreement with previously published results.

537.311.33:[546.28+546.289 1803
Direct Measurement of the Dielectric Constants of Silicon and Germanium—W. C. Dunlap, Jr, and R. L. Watters. (*Phys. Rev.*, vol. 92, pp. 1396–1397; Dec. 15, 1953.) Single-crystal Si and Ge samples were doped with gold to produce material of sufficiently high resistivity at the working temperature of 77 degrees K; and were formed into small parallel-plate capacitors. The permittivity of Si is 11.7 ± 0.2 at 1 mc and remains constant within ±1 per cent over the range 500 c/s—30 mc. The corresponding value for Ge is 15.8 ± 0.2; its much greater apparent variation with frequency is ascribed to sample inhomogeneity.

537.311.33:[546.28+546.289]:532.2 1804
Shapes of Floating Liquid Zones between Solid Rods—P. H. Keck, M. Green, and M. L. Polk. (*Jour. Appl. Phys.*, vol. 24, pp. 1479–1481; Dec., 1953.) The method described by Keck and Golay (*Phys. Rev.*, vol. 89, p. 1297; March 15, 1953.) for crystallizing Si from a melt involves holding the molten Si between two vertically aligned solid rods of Si. The shape and stability of the molten Si are investigated theoretically.

537.311.33:546.28 1805
Gold as a Donor in Silicon—E. A. Taft and F. H. Horn. (*Phys. Rev.*, vol. 93, p. 64. Jan. 1 1954.) The experimentally determined resistivity/temperature characteristic indicates that Au produces a donor level 0.33 ev above the occupied band in Si.

537.311.33:546.289 1806
Lattice-Scattering Mobility in Germanium—F. J. Morin. (*Phys. Rev.*, vol. 93, pp. 62–63, Jan. 1, 1954.) The temperature dependence of the lattice-scattering mobility of holes and electrons was determined from conductivity measurements. The variation with temperature of the ratio Hall-mobility/conductivity-mobility for holes suggests that the valence band is composed of multiple surfaces of minimum energy.

537.311.33:546.289 1807
Preparation of *p-n* Junctions by Surface Melting—K. Lehovec and E. Belmont. (*Jour. Appl. Phys.*, vol. 24, pp. 1482–1484; Dec., 1953.) A rf-heating method of preparing single or multiple *p-n* junctions is described. The influence of (*a*) heater temperature, (*b*) forced cooling, and (*c*) thickness of the slice of the

single-crystal semiconductor on the position of the solid/liquid interface is calculated from a one-dimensional model.

537.311.33:546.289:535.215 1808
Infrared Photoconductivity due to Neutral Impurities in Germanium—E. Burstein, J. W. Davisson, E. E. Bell, W. J. Turner, and H. G. Lipson. (*Phys. Rev.*, vol. 93, pp. 65–68; Jan. 1, 1954.) Experimental determination at liquid-He temperatures of the photoconductivity in the wavelength range 1–38 μ of n-type and p-type Ge containing various donor and acceptor impurities.

537.311.33:546.41-31 1809
Calcium Oxide, an Amphoteric Semiconductor—K. Hauffe and G. Tränckler. (*Z. Phys.*, vol. 136, pp. 166–178; Nov. 17, 1953.) Conductivity measurements on CaO at 600 degrees C. indicate that at oxygen pressures $<10^{-2}$ Torr CaO is a p-type semiconductor, at pressures $<10^{-2}$ Torr an n-type semiconductor. Addition of Li_2O results in p-type conduction even in a high vacuum, the conductivity increasing with increasing O_2 pressure and decreasing in high vacuum. Addition of Y_2O_3 gives opposite results, and is beneficial in alkaline-earth oxide cathodes.

537.311.33:546.47-31 1810
Conductivity and Hall Effect of ZnO at Low Temperatures—S. E. Harrison. (*Phys. Rev.*, vol. 93, pp. 52–62; Jan. 1, 1954.) Results of measurements on single-crystal and polycrystalline specimens are used in assessing the adequacy of various energy-level models of ZnO.

537.311.33:546.482.21 1811
Demonstration of Traps in CdS Single Crystals—W. Muscheid. (*Ann. Phys., Lpz.*, vol. 13, pp. 322–326; Dec. 15, 1953.) Results of an experimental investigation of the effect of illumination and rate of heating or cooling on the dark current indicate that traps are produced by incorporating oxygen in the CdS crystals.

537.311.33:546.482.21:535.215 1812
The Effect of Oxygen on the Conductivity of CdS Single Crystals—W. Muscheid. (*Ann. Phys., Lpz.*, vol. 13, pp. 305–321; Dec. 15, 1953.) Measurements are reported of the dark conductivity and photoconductivity at various pressures of air and oxygen, in the temperature range 300 degrees–700 degrees K.

537.311.33:546.482.21:538.632 1813
Hall-Effect Measurements on CdS Crystals—H. Diedrich. (*Ann. Phys., Lpz.*, vol. 13, pp. 349–352; Dec. 15, 1953.) Results of measurements in the absorption region at various illumination intensities give excess-electron mobilities of 20–400 cm/s per V/cm. The mobility depends on the conduction-electron concentration.

537.311.33:546.482.21.03 1814
Optical and Electrical Properties of Cadmium-Sulphide Single Crystals—H. Gobrecht and A. Bartschat. (*Z. Phys.*, vol. 136, pp. 224–233; Nov. 17, 1953.)

537.311.33:546.561-31 1815
Electrical Conductivity of Cuprous Oxide—A. I. Andrievski, V. I. Voloshchenko, and M. T. Mishchenko. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 90, pp. 521–523; June 1, 1953. In Russian.) Results of an experimental investigation indicate that the conductivity of the Cu_2O surface layer is proportional to the number of crystalline grains per unit area and is independent of the method of production.

537.311.33:546.682.86 1816
A Note on the Semiconducting Compound InSb—F. A. Cunnell, E. W. Saker and J. T. Edmond. (*Proc. Phys. Soc.*, vol. 66, pp. 1115–1116; Dec. 1, 1953.) Measurements were made on an ingot purified by zone melting. The variations of the conductivity and of the Hall constant with distance along the ingot from the pure end, and with temperature, are shown graphically. The results are discussed briefly.

537.311.33+538.63]:546.762.21 1817
Electrical and Galvanomagnetic Properties of Sulphides of Chromium—N. P. Grazhdankina and I. G. Fakidov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 93, pp. 429–430; Nov. 21, 1953. In Russian.) Classification according to resistivity places these sulphides intermediate between metals and semiconductors. Measurements were made of (*a*) the resistance at very low temperatures, (*b*) the Hall effect, and (*c*) the magnetoresistance effect. The results of (*a*) and (*c*) are shown graphically.

537.311.33:546.817.221 1818
Interpretation of Hall Effect and Resistivity Data in PbS and Similar Binary Compound Semiconductors—W. W. Scanlon. (*Phys. Rev.*, vol. 92, pp. 1573–1575; Dec. 15, 1953.) Discrepancies in published data on the width of the forbidden energy gap for PbS are accounted for by the occurrence of thermal changes in composition during experimental runs. Measurements confined to temperatures below 500 degrees K. for which such changes do not take place, give a figure of 0.37 ± 0.01 ev, in agreement with optical data.

537.311.33:621.314.634 1819
Mode of Operation of CdSe Intermediate Layers in Selenium Rectifiers—A. Hoffmann and F. Rose. (*Z. Phys.*, vol. 136, pp. 152–165; Nov. 17, 1953.) An experimental investigation. Results indicate that the blocking action in commercial Se rectifiers with counter electrodes containing Cd occurs at the CdSe/Se boundary, CdSe and Se being n-type and p-type semiconductors respectively. The CdSe layer has not a sufficiently high resistivity to act as an "isolating" hole layer with multipoint contact rectification.

537.312.8 1820
Effect of Composition on Change of Electrical Resistance of Ni-Mn Alloys in a Magnetic Field—I. Ya. Dekhtyar. (*Compt. Rend. Acad. Sci., U.R.S.S.*, vol. 93, pp. 637–639; Dec. 1, 1953. In Russian.) The variation of the fractional change of resistance of Ni-Mn alloys with variation of the percentage of Mn in the alloy was determined experimentally at several values of the magnetic field strength. The influence of Cu or Sn impurity atoms was also investigated. The results are presented graphically.

538.213:621.3.042.15 1821
Effective Permeabilities of Regular Configurations of Spherical Ferromagnetic Particles—I. Lucas. (*Frequenz*, vol. 7, pp. 289–295; Oct., 1953.) Estimation of effective permeability is based on an approximation to the magnetic-field distribution near the points of contact of the particles. For the closest packing, the effective permeability, μ_w, is related to the bulk permeability of the material, μ, by the formula $\mu_w=\sqrt{2}\pi \log \mu$. Agreement with measurements on sendust (Fe-Si-Al alloy) and carbonyl-iron powder cores is satisfactory.

538.221 1822
An Amplitude- and Temperature-Dependent After Effect in α-Fe at -70 degrees C—G. Sorger. (*Z. angew. Phys.*, vol. 5, pp. 406–413; Nov., 1953.) An anomaly in the complex permeability observed at about -70 degrees in Fe and Fe-Si alloys with α-type lattice was investigated. The after-effect decreases with an increase of applied field strength. The activation energy is 0.3 ev/atom. See also 1631 of 1952 (Feldtkeller et al.).

538.221 1823
Developments in Sintered Magnetic Materials—J. L. Salpeter. (PROC. I.R.E., vol. 42, pp. 514–526; March, 1954.) Reprint. See 3334 of 1953.

538.221:538.561.029.6 1824
Ferromagnetic Resonance using Waves from a Mass Emitter—K. A. Volkova. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 89, pp. 655–658; April 1, 1953. In Russian.) Measurements were made of the heat dissipated in Fe and in Ni specimens subjected to a steady magnetic field of 100–13,500 oersted and a superimposed hf field of mean frequency variable between 21.4 and 125 kmc. The latter field was produced by the wide-band mass emitter described by Glagoleva-Arkad'eva (2106 of 1943). The experimental results are shown graphically. Ferromagnetic dispersion is briefly discussed.

538.221:538.632 1825
Spontaneous Hall Effect in Ferromagnetics—J. Smit and J. Volger. (*Phys. Rev.*, vol. 92, pp. 1576–1577; Dec. 15, 1953.) Results of measurements on several specimens of Ni and Ni alloys are tabulated and discussed.

538.221:621.318.1:621.314.22.018.75 1826
Pulse Permeability and Losses in Magnetic Materials subjected to Rectangular D.C. Pulses—T. Einsele. (*Frequenz*, vol. 7, pp. 281–289; Oct., 1953.) Measurements of maximum pulse permeabilities and total losses under steady operating conditions are reported for cores of siferrit 2000 Tr. 7, permenorm 3601 and mumetal. When simplifying assumptions are made, the maximum pulse power that can be dealt with by a transformer of specified core material, for given pulse duration and repetition rate, can be calculated.

538.221:[621.318.124+621.318.134 1827
Ferromagnetic Ferrites—Y. Lescroël. (*Câbles & Transm.*, vol. 7, pp. 273–292; Oct., 1953.) Characteristic properties of ferrites are examined with reference to three materials developed since 1950, namely (*a*) fermalite, a mixed Mn-Zn ferrite, of which three principal types are manufactured; Type 1002 for coils operating at frequencies up to 1 mc; Type 2001 for wide-band transformers, and reaction coils; Type 3001 for hf power and pulse transformers; (*b*) fernilite, a Ni-Zn ferrite for applications at frequencies up to 100 mc; (*c*) fercolite, a Co ferrite used for light-weight permanent magnets.

538.221:621.318.124+621.318.134 1828
Magnetic Resonance in Ferrimagnetics—R. K. Wangsness. (*Phys. Rev.*, vol. 93, pp. 68–71; Jan. 1, 1954.) The general expression for the magnetic resonance frequency of the two-sublattice model of a ferrimagnetic crystal has essentially the same form as that originally obtained for the ferromagnetic case provided that the product of the molecular field coefficient and the net magnetization is large compared to the applied and anisotropy fields.

538.221:621.318.134 1829
Effect of Cross-Section Area and Compression upon the Relaxation in Permeability for Toroidal Samples of Ferrites—R. E. Alley, Jr., and F. J. Schnettler. (*Jour. Appl. Phys.*, vol. 24, pp. 1524–1525; Dec., 1953.) The real and the imaginary components of permeability of MnZn and NiZn ferrites were determined experimentally over the frequency range 25 kc–2 mc. Only the MnZn showed a change in the relaxation frequency with change of cross-section. The two ferrites also responded differently to compression. The results are shown graphically.

538.221:621.318.134 1830
Magnetization in Nickel Ferrite-Aluminates and Nickel Ferrite-Gallates—L. R. Maxwell and S. J. Pickart. (*Phys. Rev.*, vol. 92, pp. 1120–1126; Dec. 1, 1953.) Measurements were made on materials prepared by substituting trivalent Al and Ga for part or all of the trivalent Fe in $NiO\cdot Fe_2O_3$. Graphs show the variation with composition of the unit cell size and of the saturation magnetization extrapolated to 0 degrees K.

538.221:621.318.134 1831
Ionic Distribution deduced from the g-Factor of a Ferrimagnetic Spinel: Ti^{4+} in Fourfold Co-ordination—E. W. Gorter. (*Nature (London)*, vol. 173, pp. 123–124; Jan. 16, 1954.)

538.221:621.318.134 1832
Ferromagnetic Resonance of Nickel-Zinc Ferrites—E. I. Kondorski and N. A. Smol'kov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 93, pp. 237–240; Nov. 11, 1953. In Russian.) The relaxation times and *g*-factors of three ferrites were calculated from results of an experimental determination of the real and imaginary components of the magnetic permeability and the dielectric constant at wavelengths of 3.2 and 8.6 cm with the sample placed in a steady magnetic field. The experimental method is described. Results are tabulated and, for one ferrite, shown graphically.

538.222/.224 1833
Resonance Absorption in Metals at Centimetre Wavelengths—S. G. Salikhov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 93, pp. 241–244; Nov. 11, 1953.) Measurements are reported on 28 pure metals at temperatures of 90 degrees and 290 degrees K. using a frequency of 9.378 kmc. The table of results gives the atomic susceptibility, paramagnetic absorption, *g*-factor, line width and line intensity. Some results are also shown graphically.

538.567:539.234:546.57 1834
The Electrical Properties of Thin Evaporated Silver Films at 3000 Mc/s—F. J. Tischer. (*Z. angew Phys.*, vol. 5, pp. 413–415; Nov., 1953.) The films discussed are of a type used in microwave attenuators and terminating resistors. Radiation penetration depth and reflection and transmission coefficients are investigated, taking into account the effect of the mica base and the displacement current in the layer. The theoretical and experimental values of the reflection and transmission coefficients are compared and the material constants are calculated.

538.653.2:[546.72+546.74 1835
Effect of Plastic Deformation on the Magnetization Curves of Iron and Nickel in High-Strength Magnetic Fields—V. V. Parfenov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 93, pp. 435–438; Nov. 21, 1953. In Russian.) Report of experimental investigation using field strengths up to 10,000 oersted.

547.476.3 1836
Dielectric Properties of Some Double Tartrates—F. Jona and R. Pepinsky. (*Phys. Rev.*, vol. 92, p. 1577; Dec. 15, 1953.) No dielectric anomaly is found for the Na-Rb salt in the temperature range 4.2 degrees K.–333 degrees K. but a dielectric anomaly appears in the $NaNH_4$ salt at 109 degrees K.

548.0:539.378.3 1837
Some Predicted Effects of Temperature Gradients on Diffusion in Crystals—A. D. LeClaire, J. A. Brinkman, and W. Shockley. (*Phys. Rev.*, vol. 93, pp. 344–346; Jan. 15, 1954.) Discussion on 1122 of April and author's reply.

549.514.51:539.31 1838
Anelasticity of Quartz—R. K. Cook and R. G. Breckenridge. (*Phys. Rev.*, vol. 92, pp. 1419–1423; Dec. 15, 1953.) Measurements were based on the method of Cook and Weissler (650 of 1951). Anelasticity, i.e. internal friction, has a maximum value at a temperature between room temperature and the quartz inversion temperature of 573 degrees C., the value being greatly increased when foreign atoms are present in the lattice. Activation energies and relaxation time constants were deduced for 4 quartz bars studied.

621.315.61 1839
Progress in Dielectrics—S. Whitehead. (*Elec. Rev.* (*London*), vol. 153, pp. 1309–1313; Dec. 11, 1953.) A survey based on a paper presented at the British Association.

621.315.61:538.569.3 1840
Absorption of Millimeter Waves in Dielectric Solids—W. L. Brooks, J. H. Greig, C. Pine, W. G. Zoellner, and J. H. Rohrbaugh. (*Jour. Opt. Soc. Amer.*, vol. 43, pp. 1191–1194; Dec., 1953.) The real and imaginary parts of the complex refractive index of polystyrene, polymethyl methacrylate, and a high-melting-point paraffin were determined at wavelengths between 1.79 and 4.18 mm from measurements of the magnitudes and phase angles of the transmission coefficients, using harmonics of a 12.5 mm-λ Type-3J31 magnetron.

621.315.612.6.011.5.029.42 1841
The Dielectric Relaxation of Glass and the Pseudo-capacity of Metal-to-Glass Interfaces, Measured at Extremely Low Frequencies—J. Volger, J. M. Stevels, and C. van Amerongen. (*Philips Res. Rep.*, vol. 8, pp. 452–470; Dec., 1953.) The dispersion of the dielectric constant was determined experimentally over the frequency range 0.1–20 cps, at various temperatures. From theoretical considerations, the main relaxation time is inversely proportional to the transition probability of Na^+ ions jumping between adjacent interstices. The very high frequency- and temperature-dependent electrode capacitances are discussed.

669-426.2 1842
The Production of Fine Wires by Electrolytic Polishing—H. R. Haines and B. W. Mott. (*Jour. Sci. Instr.*, vol. 30, pp. 459–460; Dec., 1953.) A method suitable for reducing the diameter of brittle or easily oxidized metal wires is described; it has been used successfully for Th, U, Zr and nichrome.

MATHEMATICS

517.51/.52 1843
On the Summation of Infinite Series in Closed Form—A. D. Wheelon. (*Jour. Appl. Phys.*, vol. 25, pp. 113–118; Jan., 1954.)

517.942.9 1844
The Meaning of the Vector Laplacian—P. Moon and D. E. Spencer. (*Jour. Franklin Inst.*, vol. 256, pp. 551–558; Dec., 1953.) The distinction between the vector and scalar Laplace operators is discussed. A general equation is developed for the vector Laplace operator in any orthogonal curvilinear co-ordinate system; use of this equation enables electrodynamic problems to be formulated simply by means of the vector Helmholtz equation.

517.949.8 1845
On Mildly Nonlinear Partial Difference Equations of Elliptic Type—L. Bers. (*Jour. Res. Nat. Bur. Stand.*, vol. 51, pp. 229–236; Nov., 1953.) Justification is given for the use of the finite-difference method for the numerical treatment of these equations.

519.271.3:620.113.2 1846
The Efficiency of Sequential Sampling for Attributes: Part 2—Practical Applications—H. C. Hamaker. (*Philips Res. Rep.*, vol. 8, pp. 427–433; Dec., 1953.) The simplified equations valid under Poisson conditions are applied to the design of a slide rule from which the operating characteristic can be read off when two fundamental parameters are known. Part 1: 2715 of 1953.

519.272.119 1847
Error due to Finite Integration Time-Limits in the Determination of Autocorrelation Functions—P. Blassel. (*Ann. Télécommun.*, vol. 8, pp. 406–414; Dec., 1953.)

MEASUREMENTS AND TEST GEAR

529.786 1848
Annual Fluctuations in Quartz Clock Error and Frequency Drift—H. J. M. Abraham. (*Nature* (*London*), vol. 173, pp. 73–74; Jan. 9, 1954.) Report of an investigation of the performance of several quartz clocks in the southern hemisphere.

538.652.08 1849
A Method of Measuring Magnetostriction—A. W. Cochardt. (*Jour. Appl. Phys.*, vol. 25, pp. 91–95; Jan., 1954.) The absolute magnitude of magnetostriction is determined from torsion tests on unmagnetized and magnetized wires. The error is about 3 per cent at stresses >1000 psi.

621.317.3:621-526 1850
Measurement of L.F. Response of Servo-Mechanism Components—E. J. P. Long. (*Engineer* (*London*), vol. 196, pp. 722–725; Dec. 4, 1953.) A description is given of apparatus for producing a sinusoidal signal in the frequency range 0.5–20 cps, together with suitable equipment for measuring attenuation and phase shift at these frequencies. The method consists in operating a magslip so that one complete revolution of the rotor corresponds to one cycle of the required low frequency, and using the magslip as a mechanical modulator for a hf carrier. The display end of the apparatus consists of a two-channel oscilloscope.

621.317.3:621.385.029.6 1851
Rapid Method of Testing Magnetrons in the Nonoperating State—J. W. Dodds. (*Le Vide*, vol. 8, pp. 1429–1431; Nov., 1953.) The resonance frequency and the effect of loading are determined using an arrangement in which the magnetron terminates a line fed by a frequency-modulated klystron. The signal is picked up by a standing-wave detector, and the resonance condition is indicated by the shape of the curve displayed on a cro. Under-coupling and over-coupling are easily identifiable. Relevant theory is given.

621.317.311 1852
The Measurement of Very Small Direct Currents—M. W. Jervis. (*Electronic Eng.*, vol. 26, pp. 100–105; March, 1954.) A review of methods used for measuring currents in the range $10^{-16}-10^{-6}$A, with particular reference to thermionic and capacitor modulator electrometers. 32 references.

621.317.329:621.373.413 1853
Simple Method of Measurement of the Distribution of the Electromagnetic Field in a Resonant Cavity—A. Septier. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 658–660; Feb. 8, 1954.) The determination of the field is based on observed variations of resonance frequency with displacement of a suitably shaped obstacle inside the cavity. Frequency variations of 1 part in 10^5 can be detected.

621.317.335.3:546.171.1 1854
Dispersion in Ammonia in 3 cm Region—Krishnaji and P. Swarup. (*Z. Phys.*, vol. 136, pp. 374–378; Dec. 8, 1953. In English.) The variation of electric susceptibility with pressure was determined experimentally using a standing-wave technique; the results are tabulated and shown graphically. The value of the susceptibility at atmospheric pressure is 5.3 $\times 10^{-3}$.

621.317.342 1855
Measurements of Phase Angles—A. van Weel. (*Philips Res. Rep.*, vol. 8, pp. 471–475; Dec., 1953.) Discussion of an improved system based on the method described previously (3659 of 1953).

621.317.382:538.632:537.311.33 1856
Application of the Hall Effect in a Semiconductor to the Measurement of Power in an Electromagnetic Field—H. M. Barlow. (*Nature* (*London*), vol. 173, pp. 41–42; Jan. 2, 1954.)

621.317.443.029.4/.5 1857
A Radio-Frequency Permeameter—P. H. Haas. (*Jour. Res. Nat. Bur. Stand.*, vol. 51, pp. 221–228; Nov., 1953.) Theory and description of a permeameter for frequencies up to 20 mc. The measurement of permeability and losses is made by inserting the ferromagnetic material in the form of a toroidal core into the short-circuited secondary of a transformer, the primary of which is connected to a rf bridge or *Q* meter.

621.317.7:537.54:621.396.822.029.64 1858
A Microwave Wide-Band Noise-Generator in Coaxial-Line Form—W. Friz. (*Fernmeldetech. Z.*, vol. 6, pp. 583–587; Dec., 1953.) A description is given of the characteristics and construction of a gas-discharge noise generator for the 6.5–20-cm wavelength range, with the discharge path inside a coaxial line. The advantages of this type over the diode type are (*a*) nearly constant internal resistance, and (*b*) constant power distribution over a wide frequency range. The upper and lower frequency limits are discussed.

621.317.733 1859
Resistance Bridge Sensitivity and Output Formulas—P. M. Andress. (*Elect. Eng.*, vol. 72, p. 1095; Dec., 1953.) Digest of paper to be published in *Trans. AIEE*, vol. 72, 1953.

621.317.733:621.314.2 1860
Bridges with Coupled Inductive Ratio Arms as Precision Instruments for the Comparison of Laboratory Standards of Resistance or Capacitance—C. W. Oatley and J. G. Yates. (*Proc. IEE*, part III, vol. 101, pp. 91–100; March, 1954.) An estimate is made of errors likely to be caused by deviation of the transformer from the ideal, in respect of imperfect coupling, winding resistance, stray capacitance and eddy currents in the core. Measurements on six transformers designed to give low errors are reported. For ratios up to 100:1 there is no difficulty in constructing transformers whose effective ratios are equal to their turns ratios within 1 part in 10^4; even at 1000:1 the error need not exceed 1 or 2 parts in 10^4.

621.317.733:621.372.2:621.315.212 1861
Some Steady-State Characteristics of Short Irregular Lines—Rosen. (*See* 1668.)

621.317.755 1862
A Portable High-Speed Cathode-Ray Oscillograph—S. Waring and B. Murphy. (*Jour. Sci. Instr.*, vol. 30, pp. 469–471; Dec., 1953.) The instrument described uses a sealed-off cr tube and an asymmetrical thyratron timebase which can be operated to give either a single sweep or a repeated sweep at frequencies up to 500 cps.

621.317.761 1863
A Stroboscopic Frequency Meter—C. W. McLeish and D. H. Rumble. (PROC. I.R.E., vol. 42, pp. 594–596; March, 1954.) The instrument described is capable of giving a direct indication to the nearest kilocycle in the range 3–30 mc, and is adapted for building in with the signal source (e.g. receiver local oscillator).

621.317.784.029.64 1864
Automatic Milliwattmeter for Electromagnetic Radiation—J. C. van den Bosch and F. Bruin. (*Appl. Sci. Res.*, vol. B3, pp. 260–264; 1953.) An instrument for the 1–10-cm waveband, with two ranges giving respectively 1 and 10 mw full scale deflection, comprises a Wheatstone bridge, one arm of which is formed by a thermistor enclosed in a waveguide; the bridge is balanced by varying the current through the thermistor by means of the tube circuit described.

621.317.789:621.373.2.029.65/.66 1865
Boltzmann Interferometer—J. L. Farrands and J. Brown. (*Wireless Eng.*, vol. 31, pp. 81–83; April, 1954.) Use of the interferometer for investigating the power spectrum of spark-type oscillators is described. The spectrum is given by the Fourier transform of the interference pattern.

621.373.029.3:621.376.222 1866
A Low-Distortion Pulse Modulator—F. Brunner. (*Öst. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 7, pp. 147–149; Nov./Dec., 1953.) Description of a circuit designed for investigating transients in electroacoustic systems. A signal tone is modulated by a square-wave voltage the frequency of which is continuously variable between 4 cps and 100 kc. The modulator circuit comprises two hexodes so arranged that the modulating voltage itself is eliminated in the output.

621.373.1.029.42:621-526 1867
Forcing Function Generator using Conductive Plastic—L. W. Norman. (*Elect. Eng.*, vol. 72, p. 1112; Dec., 1953.) Digest of paper to be published in *Trans. AIEE*, vol. 72, 1953. A potentiometer made of conductive plastic is used to obtain a signal of frequency between 0.05 and 60 cps and of controllable waveform, for testing servomechanisms.

621.397.2.001.4 1868
The Application of Pulse Technique in the Testing of Television Transmission Circuits—H. Röschlau. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 187–190; 1953.) The pulse-echo meter used by the Federal German Post Office provides either direct-voltage square-wave pulses of 12.5×10^{-8} sec duration, or alternately positive and negative square-wave pulses of duration 4.7×10^{-8} sec. It can be used either for the detection of cable irregularities or, with the addition of a special filter, as mismatch indicator. The double-reflection test set operates with a sine-squared test pulse having a repetition rate equal to the television-picture line frequency.

535.824:621.397.6 1869
Television Optics—Zschau. (*See* 1932.)

621.317.083.7 1870
Telemetering Systems—W. Bösch. (*Microtecnic*, vol. 6, pp. 303–311 and 1953; 1952. vol. 7, pp. 19–24, 85–91, 142–148 and 256–260.) A comprehensive review.

621.384.611 1871
Acceleration of Partially Stripped Heavy Ions—R. S. Livingston. (*Nature* (*London*), vol. 173, pp. 54–57; Jan. 9, 1954.) Description of the 63-inch cyclotron at Oak Ridge (Tennessee) and of nuclear research carried out with this machine; the hot-cathode source used to produce partially stripped ions is a unique feature.

621.384.611 1872
Characteristics of a Proposed Double-Mode Cyclotron—M. J. Jakobson and F. H. Schmidt. (*Phys. Rev.*, vol. 93, pp. 303–305; Jan. 15, 1954.) Operation of a two-dee fixed-frequency cyclotron with the dee voltages either in phase opposition or in the same phase is discussed. The system would be useful for accelerating ions heavier than α particles.

621.384.612 1873
The Profile of the [pole] Pieces of Strong-focusing Cosmotrons with Small Diameter—G. Sasson. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 885–888; Feb. 22, 1954.)

621.384.622 1874
Improvement of the Performance of a Linear Accelerator by Bunching the Electrons before Injection—M. Papoular. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 789–791; Feb. 15, 1954.) Analyses indicates that considerable increase of efficiency is attainable by this method.

621.384.622.1 1875
The Magnitude of the Divergence Caused by the Accelerating Gaps in Linear Ion Accelerators—M. Y. Bernard. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 675–677; Feb. 8, 1954.)

621.385.833 1876
A New Microscope Principle—J. M. Cowley. (*Proc. Phys. Soc.*, vol. 66, pp. 1096–1100; Dec. 1, 1953.) A high-resolution image may be derived from a large number of "dark field" images of normal resolution obtained by varying the angle of incidence of the electron beam in a standard electron microscope. The practical limitations of this theoretical result are discussed.

621.385.833 1877
Calculation of the Axial Potential of Electron Lenses Constituted by Two Coaxial Cylinders of Different Diameters—P. Ehinger. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 328, pp. 879–881; Feb. 22, 1954.) The axial position of the equipotential surface whose mean curvature is zero is introduced as reference parameter.

621.387.4:519.2 1878
Stochastic Theory of Electronic Counters—F. Pollaczek. (*Compt. Rend. Acad. Sci* (*Paris*), vol. 238, pp. 766–768; Feb. 15, 1954.)

621.387.424 1879
The Behaviour of Counters with External Cathode and Pure Methyl-Alcohol Filling, Subjected to Gamma Radiation—D. Blanc. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 673–675; Feb. 8, 1954.) Report of an experimental investigation to determine the value of pressure for best performance.

621.387.424 1880
The Delay in the Build-up of Halogen-Quenched Counters—D. van Zoonen. (*Appl. Sci. Res.*, vol. B3, pp. 377–389; 1953.) Continuation of the experiments of van Zoonen and Prast (1463 of 1953).

621.389:620.16 1881
Symposium on "Vibration Methods of Testing"—(*Jour. Brit. IRE*, vol. 14, pp. 93–126; March, 1954.) Text of, and discussion on, four papers, as listed below, dealing with the application of electronics to the measurement of mechanical vibrations.
Vibration Generators, Ancillary Equipment and Applications (pp. 94–100—H. Moore.
Electronic Stroboscopes (pp. 101–105)—F. M. Savage.
Resistance Strain Gauges and Vibration Measurement (pp. 106–114)—P. Jackson.
Electronic Aids to Vibration Measurement (pp. 115–124)—R. K. Vinycomb.

621.397.9:612.1 1882
Particle Counting by Television Techniques—L. E. Flory and W. S. Pike. (*RCA Rev.*, vol. 14, pp. 546–556; Dec., 1953.) Description, with circuit diagram, of a closed-circuit television system designed for counting blood cells.

621.385 1883
Electron Optics. [Book Review]—Klemperer. (*See* 1976.)

PROPAGATION OF WAVES

621.396.11 1884
Radio Communication by Scattering from Meteroic Ionization—V. R. Eshleman and L. A. Manning. (PROC. I.R.E., vol. 42, pp. 530–536; March, 1954.) The amplitude and duration of forward scattered echoes from individual meteor trails, and the probability of detecting randomly oriented trails over an oblique propagation path are examined. The results indicate that for a frequency of about 15 mc the trail ionization would give a practically continuous signal over a path of about 1000 km. For vhf signals, scattering from the meteor trails is at least an important contributory factor in the propagation; further observations are required in order to evaluate this factor.

621.396.11:551.510.52 1885
The Troposphere as a Medium for the Propagation of Radio Waves: Part 2—H. Bremmer. (*Philips Tech. Rev.*, vol. 15, pp. 175–181; Dec., 1953.) Continuation of survey noted in 1548 of May. A discussion of scattering is presented, based on theory and observations.

621.396.11:551.510.535 1886
Real and Complex Wave Polarization in the Ionosphere—J. C. W. Scott. (*Jour. Geophys. Res.*, vol. 58, pp. 437–443; Dec., 1953.) "The relation between the polarization ellipse in the wave front and the complex polarization at vertical incidence in a slowly-varying horizon-

tally-stratified ionosphere is reviewed. Charts and a table are given showing the sense, orientation and eccentricity of the polarization ellipse under all conditions of plasma frequency, collisional frequency, wave frequency, and magnetic field intensity and direction."

621.396.11:551.510.535 1887

On the Coupled Wave Equations of Magneto-ionic Theory—J. M. Kelso. (*Jour. Geophys. Res.*, vol. 58, pp. 431–436; Dec., 1953.) The coupled wave equations describing the propagation of a radio wave incident vertically on a horizontally stratified ionosphere [1884 of 1942 (Försterling)] and the basic defining relations are presented so as to show explicitly the confusing effects resulting from the different forms used by other authors. The various notations are tabulated.

621.396.11:551.510.535 1888

High-Frequency Scatter-Sounding Experiments at the National Bureau of Standards—R. Silberstein. (*Science*, vol. 118, pp. 759–763; Dec. 25, 1953.) Four basic types of presentation used in investigating skip distance by means of back-scattered echoes are discussed, namely (*a*) intensity/range at fixed frequency, (*b*) range/time at fixed frequency, (*c*) ppi, and (*d*) sweep-frequency records. Experimental results confirm that skip-distance determination by back-scatter methods is not always possible. The technique calls for skilled personnel familiar with the regular behavior of the ionosphere in the region.

621.396.11:551.510.535 1889

Maximum Usable Frequencies and Lowest Usable Frequencies for the Path Washington to Resolute Bay—G. H. Hanson, H. V. Serson, and W. Campbell. (*Jour. Geophys. Res.*, vol. 58, pp. 487–491; Dec., 1953.) WWV transmissions were monitored continuously at Resolute Bay, 4009 km away, for a period of one year; values of the muf's and luf's were determined from the records. The predicted muf's were found to be too low during the night and, for some months, too high during the day. The luf's were lower than expected for propagation by one reflection from the F_2 layer.

621.396.11.029.55 1890

Measurements of Angle of Arrival in the Short-Wave Range—K. Vogt. (*Fernmeldetech. Z.*, vol. 6, pp. 537–539; Nov., 1953.) The procedure and results of measurements reported since 1934 are reviewed. The need for further investigation of the dependence of angle of arrival on sunspot cycle and direction of propagation is evident.

621.396.11.029.6 1891

Metre, Decimetre and Centimetre Waves Far beyond the Horizon—E. Roessler. (*Frequenz*, vol. 7, pp. 313–319; Nov., 1953.) A review of recent literature on long-distance uhf propagation, and of the consequences for the planning of radio transmissions and links.

621.396.11.029.62 1892

A Simple Method for estimating U.S.W. Propagation within Line of Sight—E. Prokott. (*Telefunken Ztg.*, vol. 26, pp. 346–352; Dec., 1953.) The distribution of illumination intensity over a relief map illuminated by a lamp placed at the top of a model of the transmitter antenna mast is in fair agreement with measured line-of-sight usw field-strength distribution. A correction factor for diffraction is given.

621.396.11.029.62 1893

Report of Propagation Tests on 45 and 66.6 Mc/s in the Central Mediterranean—J. Roux. (*Ann. Radioélect.*, vol. 8, pp. 318–330; Oct., 1953.) During summer 1952 and early 1953 continuous records were made of field strength, signal/noise ratio, interference, and meteorological data for two over-sea transmission paths of 245 and 285 km respectively, proposed for radio relay links between Europe and North Africa. Result sof an analysis of the records are presented; these include details of interfering signals, the most remarkable of which were from Berlin, from Monte Cavo, near Rome, and television transmissions from Paris.

621.396.11.029.64/.65 1894

Atmospheric Attenuation of Millimetre and Centimetre Waves—A. Perlat and J. Voge. (*Ann. Télécommun.*, vol. 8, pp. 395–405; Dec., 1953.) Tables and curves showing the attenuation per kilometer in the 10-mm-10-cm frequency range due to various forms of precipitation, fogs and cloud are assembled from the literature. Meteorological statistics for France and North Africa are also presented. From these, the maximum attenuation to be expected over a given path during a particular percentage of time can be estimated for each meteorological factor.

RECEPTION

621.376.33 1895

High-Linearity Demodulation of Frequency-Modulated Oscillations—H. Meinke. (*Fernmeldetech. Z.*, vol. 6, pp. 571–577; Dec., 1953.) The conditions are derived for high linearity of a demodulator consisting of four reactances, one effective resistance and one rectifier. The characteristics of several possible circuits are considered briefly; one circuit is considered in detail.

621.396.621 1896

Analysis of a Limiter as a Variable-Gain Device—L. R. Kahn. (*Elect. Eng.*, vol. 72, pp. 1106–1109; Dec., 1953.) The limiter is treated as a device whose gain is varied so that the amplitude-modulation component of the input signal is eliminated. The method is demonstrated by considering the side-band spectrum of the phase-modulation component of a two-tone signal and the mechanism of the common-limiter diversity system.

621.396.621 1897

Post-War Development of Broadcast Receivers [in Western Germany]—E. Klotz and G. Schaffstein. (*Telefunken Ztg.*, vol. 26, pp. 372–382; Dec., 1953.) A survey with particular reference to the development of vhf FM receivers.

621.396.621.54:621.314.7 1898

The Transistor as a Mixer—J. Zawels. (Proc. I.R.E., vol. 42, pp. 542–548; March, 1954.) Operation is considered for signal frequencies up to vhf, three frequency ranges being distinguished, namely (*a*) the lowest range, in which the conversion performance depends primarily on the α value of the transistor; (*b*) a middle range in which the base resistance is the dominant factor; (*c*) the high-frequency range where the capacitance shunting the reverse-biased emitter determines the performance. Methods for calculating the conversion gain are indicated, and some measurements are reported. At the low frequencies a junction transistor can give as good a performance as a crystal-diode mixer followed by a junction transistor amplifier. Point-contact types can give conversion gain in the vhf band, but their noise factor is relatively high.

621.396.82:621.376.3 1899

Reception of an F.M. Signal in the Presence of a Stronger Signal in the Same Frequency Band, and Other Associated Results—R. M. Wilmotte. (*Proc. IEE*, part III, vol. 101, pp. 69–75; March, 1954.) Analysis based on the amplitude variation of the vector resultant of two FM signals of different frequencies is used to derive methods for (*a*) receiving a weak signal in the presence of a stronger one, (*b*) reducing the interfering effect of a weak signal on a stronger one, and (*c*) transmitting a number of weak coded messages with an FM signal and in the same band. The bandwidths required by various FM multiplex systems are compared.

621.396.822 1900

On the Detection of Sure Signals in Noise—R. C. Davis. (*Jour. Appl. Phys.*, vol. 25, pp. 76–82; Jan., 1954.) Several criteria for an optimum pre-detection filter are compared. When the input noise is Gaussian, the optimum filter is identical with the linear filter which maximizes the output signal/noise ratio. The stability of the optimum filter is discussed in some detail, the discussion applying in the practical case only to finite-memory filters. The determination of an optimum signal shape of given energy content and duration is considered.

621.396.828 1901

Review of the Pilot Radio-Interference-Suppression Campaign in Iserlohn—O. Schmidt. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 191–193; 1953.) Details of the organization, measurements and equipment required in dealing with some 600 industrial and domestic sources of interference are given. Difficulties in design and supply at reasonable cost of efficient mains filter units were encountered.

621.396.621(083.72) 1902

British Standard Glossary of Terms for the Electrical Characteristics of Radio Receivers (B.S.2065:1954) [Book Review]—Publishers: British Standards Institution, London, 1954, 6s. (*Wireless World*, vol. 60, p. 188; April, 1954.)

STATIONS AND COMMUNICATION SYSTEMS

621.376.3:517.942.922 1903

The Representation of a Frequency-Modulated Oscillation by means of Bessel Functions—V. Poledna. (*Frequenz*, vol. 7, pp. 336–338; Nov., 1953.)

621.376.3:621.396.82 1904

Reception of an F.M. Signal in the Presence of a Stronger Signal in the Same Frequency Band, and Other Associated Results—Wilmotte. (*See* 1899.)

621.376.5 1905

Gaussian Pulses—J. P. Vasseur. (*Ann. Radioélect.*, vol. 8, pp. 286–300; Oct., 1953.) The advantage of using Gaussian-shaped pulses in pm systems is noted. Pulses of this shape may be derived from a narrow pulse by means of a *RC* filter chain. A band-pass filter of this type is described. The AM and FM distortion which may occur in transmitting these pulses is calculated.

621.39.001.11 1906

Bandwidth, Holding Time and Signal/Noise Ratio for Various Forms of Communication in Relation to Shannon's Theory—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 6, pp. 555–563; Dec., 1953; and vol. 7, pp. 33–43; Jan., 1954.)

621.39.001.11 1907

The Application of Information Theory to Data-Transmission Systems, and the Possible Use of Binary Coding to Increase Channel Capacity—J. F. Coales. (*Proc. IEE*, part III, vol. 101, p. 76; March, 1954.) Discussion on 253 of January.

621.39.001.11 1908

Exact Interpolation of Band-Limited Functions—A. Kohlenberg. (*Jour. Appl. Phys.*, vol. 24, pp. 1432–1436; Dec., 1953.) A general formula is derived for the spectrum of a multiply periodic, AM sequence of pulses. The result is used to show that a function which lies in a frequency band (W_0, W_0+W) is completely determined by its values at a set of points of density $2W$, the points consisting of two similar groups with spacing $1/W$ shifted with respect to each other. This verifies a supposition commonly accepted in communication theory.

621.395.724:621.395.44 1909

Standardized Equipment Type 51L—J. Malézieux, R. Sueur and M. Lebedinsky.

(*Câbles & Transm.*, vol. 7, pp. 313–324; Oct., 1953.) Description of new French Post Office equipment standardized on the rack principle, particularly for carrier-current repeater stations.

621.396:621.372.5 1910
Radio-Frequency Phase-Difference Networks: a New Approach to Polyphase Selectivity—M. G. Cifuentes and O. G. Villard, Jr. (PROC. I.R.E., vol. 42, pp. 588–593; March, 1954.) The possibilities of polyphase methods [313 of 1951 (Macdiarmid and Tucker)] for obtaining selectivity are extended by using rf instead of af phase-difference networks. The design of these networks is discussed. A polyphase selective system suitable for ssb transmission or reception is described.

621.396.1 1911
Theoretical Investigation of the Appropriate Spatial Distribution of Frequency Channels for Uniform Coverage of Large Plane Areas—P. Thiessen. (*Frequenz*, vol. 7, pp. 319–325; Nov., 1953.) Simplifying assumptions are made, so that the location of transmitting stations, when the number of available frequency channels is limited, can be considered solely from a geometrical point of view. Two types of grid-net are considered, rectangular and triangular, and geometrical patterns are reviewed for which adjacent-channel and co-channel interference shall be a minimum. Under ideal conditions the least number of channels required is 9, and the desirable number for adequate coverage without interference is 12 or 13.

621.396.5:621.318.57 1912
A Differential-Feedback Suppressor with Independent Two-Way Threshold-Level Adjustment—F. Rumpel. (*Fernmeldetech. Z.*, vol. 6, pp. 540–543; Nov., 1953.) A description of the voice-operated switching system used at Frankfurt for the transatlantic radiotelephone service. A comparator circuit controls the switching operation during pauses and break-in working. Sensitivity in each direction can be adjusted independently to prevent interruptions due to noise at the receiving end.

621.396.61/.62:621.373.421.13 1913
A Frequency-Generating System for V.H.F. Communication Equipment—G. J. Camfield. (*Proc. IEE*, part III, vol. 101, pp. 85–90; March, 1954.) Technique developed in connection with mobile communication systems is described. 2000 channel frequencies are provided, using only 32 crystals, comprising three groups of ten plus one group of two. Selection is accomplished by positioning four switches, the operation thereafter being automatic. Compared with the use of a separate crystal to control each frequency, the system offers the advantages of flexibility, quartz economy, simpler maintenance, and the possibility of spacing the chennels more closely.

621.396.65:621.396.93 1914
Some Aspects of the Design of V.H.F. Mobile Radio Systems—E. P. Fairbairn. (*Proc. IEE*, part III, vol. 101, pp. 53–64; March, 1954. Discussion, pp. 64–68.) Present practice in mobile radio systems is outlined. Methods of operation of complete simplex, two-frequency simplex ("dusimplex") and duplex systems are described. The merits of AM and FM are compared, and the possibility of making better use of available frequency bands is discussed. Valve problems are mentioned briefly.

621.396.66:621.372.56 1915
A Programme Fading Circuit—R. C. Whitehead. (*B.B.C. Quart.*, vol. 8, pp. 252–256; Winter, 1953/1954.) Attenuators comprising temperature-sensitive resistors are used, the resistance values are controlled by variation of a locally generated supply. Methods of segregating the control current from the signal current are considered; frequency filtering is recommended for this purpose.

621.396.66:621.396.712 1916
Monitoring and Two-Channel Operation of the RIAS Berlin Medium-Wave Transmitter—O. v. Broecker. (*Telefunken Ztg.*, vol. 26, pp. 342–345; Dec., 1953.) The common control equipment for the 100-kw and 200-kw transmitters is described. The operation of these transmitters either in parallel (see 1943 below) on 989 kc, or separately on 683 kc and 989 kc respectively, is also described.

621.396.7:621.396.65.029.55 1917
The Tangier Radio Relay System—C. G. Dietsch. (*RCA Rev.*, vol. 14, pp. 557–568; Dec., 1953. *Trans. I.R.E.*, vol. CS-2, pp. 65–68; Jan., 1954.) An illustrated descriptive account. The station relays telegraph, telephone and radio-photo transmissions between New York and several African, Asian and European stations. Factors considered in the choice of the geographical location are discussed. Double- and triple-diversity receivers and 1–15-kw transmitters with rhombic aerials are used, with operating frequencies in the 4–22-mc range.

621.396.71.029.51 1918
Radio Equipment of the Long-Wave Telegraphy Station at Mainflingen near Aschaffenburg—E. Meinel. (*Fernmeldetech. Z.*, vol. 6, pp. 528–537; Nov., 1953.) Illustrated descriptions are given of two 60-kw and four 50-kw transmitters operating on frequencies between 46 and 125 kc in the European telegraphic service. Specifications are given for the 100–200-m antenna masts which carry four T antennas and two triangular fan-type antennas, the latter equipped for diplex operation.

621.396.932+621.396.969.3 1919
Radio in the Service of the Seaman—C. V. Robinson. (*Proc. IEE*, part I, vol. 101, pp. 27–28; Jan., 1954.) Chairman's address, IEE Southern Center. Communications and navigational aids are reviewed.

621.396.97+621.397.5 1920
Inaugural Address [of I.E.E. President]—H. Bishop. (*Proc. IEE*, part I, vol. 101, pp. 1–10; Jan., 1954.) A review of the development of broadcasting and television, with special reference to the services of the B.B.C., and to the need for international cooperation. See also *Nature* (*London*), vol. 173, pp. 248–249; Feb., 1954.

SUBSIDIARY APPARATUS

621-526 1921
Step- to Frequency-Response Transforms for Linear Servo Systems—L. C. Ludbrook. (*Electronic Eng.*, vol. 26, pp. 27–30, 51–55, and 122–126; Jan./March, 1954.) Five known methods for finding frequency response from a given graph of linear mode step response are discussed. Theory and practical computing routines are given for a new method based on straight-line-segment approximation to the given step-response graph. Approximate laws relating cut-off frequency and frequency and amplitude of maximum response to the shape and time scale of the step response are derived. Discrepancies between experimental results and those derived from the theory are ascribed mainly to nonlinearity of practical systems.

621.3.013.783 1922
Electromagnetic Shielding with Transparent Coated Glass—E. I. Hawthorne. (PROC. I.R.E., vol. 42, pp. 548–553; March, 1954.) The shielding effectiveness of multilayer coated plane glass structures is investigated by determining their transmission factor for plane em waves. Three types of glass are considered, at frequencies up to 10 kmc. Transmission-line theory is used in the analysis, the coatings being represented by shunting resistances. Theoretical and experimental results are compared. Attenuations of 40 db and more can be obtained with tolerable loss of transparency.

621.311.6:537.311.33:539.165 1923
The Electron-Voltaic Effect in *p-n* Junctions induced by Beta-Particle Bombardment—Rappaport. (*See* 1799.)

621.311.6:621.317.3 1924
Supply for Precision Measurement—(*Elect. Jour.*, vol. 151, pp. 1927–1929; Dec. 11, 1953.) A low-impedance generator giving a highly stable comprehensive supply over the frequency range 40–2500 cps is based on a Wien-bridge oscillator with stabilized dc supplies.

621.316.722.1 1925
Voltage Stabilizers for Microwave Oscillators—F. A. Benson and G. V. G. Lusher. (*Electronic Eng.*, vol. 26, pp. 106–110; March, 1954.) Analysis indicates that the effect of variation of valve heater voltage in commonly used stabilizers [3555 of 1949 (Benson)] is not serious in the case of series-parallel arrangements but may be serious where a single series valve is used.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2.001.4 1926
The Application of Pulse Technique in the Testing of Television Transmission Circuits—Röschlau. (*See* 1868.)

621.397.311.2:621.372.55 1927
Aperture Compensation for Television Cameras—R. C. Dennison. (*RCA Rev.*, vol. 14, pp. 569–585; Dec., 1953.) Aperture distortion, caused by the finite size of the scanning aperture, can be compensated by using a dispersionless transversal filter [55 of 1941 (Kallmann)] designed to provide a given frequency and phase characteristic. Design equations for fixed- and variable-boost aperture compensators are developed.

621.397.5:621.396.1 1928
Television Coverage—J. A. Saxton. (*Wireless World*, vol. 60, pp. 173–176; April, 1954.) The influence of the terrain on the service areas of vhf and uhf transmitters is examined. Analysis of field-strength measurements for an area such as that around London indicates that the median field strength varies with distance according to a law of the same form as that for a smooth spherical earth, but the absolute measured values fall progressively below the theoretical values as the frequency increases. The significance of these results for television transmission in bands I, III, IV and V is discussed.

621.397.5+621.396.97 1929
Inaugural Address [of I.E.E. President]—Bishop. (*See* 1920.)

621.397.5:771.53 1930
Special Films for Recording Television Transmissions—W. Behrendt. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 235–237; 1953.) Details are given of an orthochromatic film whose gradation can be varied over a wide range in processing.

621.397.5:778.5 1931
Some Fundamental Aspects of Telerecording—C. B. B. Wood. (*Jour. Telev. Soc.*, vol. 7, pp. 143–151; Oct./Dec., 1953.) The advantages and disadvantages of waveform and picture recording are examined, and practical systems of the latter class are discussed.

621.397.6:535.824 1932
Television Optics—H. Zschau. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 217–223; 1953.) A survey of the applications of optics in television and of television in optics. 40 references.

621.397.61:621.372.553 1933
Design of Phase Equalizers for Television-Transmission Systems—E. Menzer and H. Voelkel. (*Fernmeldetech. Z.*, vol. 6, pp. 578–582; Dec., 1953.) The characteristics of an asymmetrical bridged-*T* section are given; application to the correction of a given phase distortion is discussed. A calculation is made for a

phase equalizer for a 5-mc low-pass filter, and the theoretical performance is compared with experimentally determined phase/frequency characteristics. A variable phase equalizer is briefly described.

621.397.611:778.5 1934
Scanning Film Negatives for Television—E. Legler. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 232–234; 1953.) Gamma correction and the production of correct black-level values are discussed and a suitable circuit is illustrated. The interference ratio obtained with negatives is smaller than with positives.

621.397.611.2 1935
Television Pick-up Tube for both Light and X-Ray Pictures—L. Heijne, P. Schagen, and H. Bruining. (*Nature* (*London*), vol. 173, p. 220; Jan. 30, 1954.) Brief note of a camera tube using a photoconductive layer of lead oxide evaporated on to a pyrex window.

621.397.611.2:621.397.82 1936
Interference Ratio and Types of Interference in Picture Scanners used in Germany at Present—W. Dillenburger. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 209–216; 1953.) A discussion of the effects on picture quality of shot noise, mains hum, microphony, and the granular structure of the mosaic target or fluorescent screen. The supericonoscope, superorthicon, and a flying-spot scanner are considered.

621.397.62 1937
The Importance of the D.C. Component—D. C. Birkinshaw. (*Jour. Telev. Soc.*, vol. 7, pp. 176–177; Oct./Dec., 1953.) Discussion on 552 of February.

621.397.62 1938
Two-Band Television Receivers—G. H. Russell. (*Wireless World*, vol. 60, pp. 189–192; April, 1954.) The choice of a suitable IF for British receivers covering bands I and III is discussed. The best frequency appears to be 35.25 mc.

621.397.62:621.396.662 1939
12-Channel Television Tuner—(*Wireless World*, vol. 60, pp. 162–164; April, 1954.) The tuner covers the vhf bands I and III, with five switch positions for band I and seven for band III. It comprises a cascade signal-frequency amplifier and a frequency changer, and provides an output at IF. A simplified circuit diagram is shown.

621.397.62:621.396.662.029.63 1940
A Capacitive-Tuned Ultra-High-Frequency Television Tuner—E. M. Hinsdale, Jr., and I. D. Baumel. (*RCA Rev.*, vol. 14, pp. 461–481; Dec., 1953.) A tuned circuit is described consisting of a split-stator capacitor and the metal box enclosing it. Two of these circuits are used in the preselector and one in the oscillator of a 470–890-mc tuner. The construction, circuit details, and performance figures are given.

621.397.621.2:621.397.335 1941
Flywheel Scanning and Synchronizing Circuits—H. Fairhurst. (*Jour. Telev. Soc.*, vol. 7, pp. 152–159; Oct./Dec., 1953.) Various systems used for flywheel synchronization are examined. Methods for eliminating effects due to variations of temperature and of mains frequency are discussed.

TRANSMISSION

621.376.32 1942
Investigation of a Reactance-Valve Circuit for the Frequency Modulation of Oscillators—H. Behling. (*Telefunken Ztg.*, vol. 26, pp. 367–370; Dec., 1953.) The circuit given is sufficiently linear and stable for use in the usw band. The modulation characteristic is derived for an arrangement using two type EF 802 tubes. A comparison is made between calculated and experimentally determined characteristics for an operating frequency of about 7.5 mc.

621.396.61:621.396.712 1943
Parallel Connection of Two Independent High-Power Sections in the RIAS Berlin 300-kW Medium-Wave Broadcasting Transmitter—K. Müller. (*Telefunken Ztg.*, vol. 26, pp. 335–341; Dec., 1953.) The bridge-type network used in coupling the output of the 100-kw and 200-kw transmitters to the common antenna is described and design considerations are discussed. Zero phase difference between the two transmitters is obtained by inserting a phase-shifting network in the line between the master oscillator and the 100-kw transmitter.

621.396.61:621.396.8 1944
Subjective and Objective Testing of the Transmission Quality of Medium-Wave Transmitters—K. H. Baer and H. Lauer. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 5, pp. 178–186; 1953.) Details of subjective tests of music transmission on four medium-power transmitters are given and the results compared with objective distortion measurements. Three transmitters gave very satisfactory performance. Making allowance for distortion in the program circuits, distortion in the transmitter should be less than 1 per cent between 120 and 1000 cps to give good musical quality but may be slightly higher outside this range.

TUBES AND THERMIONICS

537.533:621.385.029.6 1945
Space Charge Waves in Inhomogeneous Electron Beams—G. Kent. (*Jour. Appl. Phys.*, vol. 25, pp. 32–41; Jan., 1954.) A sheet beam confined between two plane electrodes by a large applied magnetic field provides the simplified model required for analysis. It is assumed that space charge is not neutralized, hence the beam is inhomogeneous in velocity and charge density. For small inhomogeneity and continuous variation of all quantities as functions of the inhomogeneity, growing waves are not possible. The divergence between this conclusion and those of Haeff (1825 of 1949) is discussed.

537.533.8 1946
The Angular Distribution of the Secondary Electrons of Soot—J. L. H. Jonker. (*Philips Res. Rep.*, vol. 8, pp. 434–440; Dec., 1953.) Experimental determination using equipment previously described (1769 of 1952). The results for soot and nickel are compared.

621.314.63+621.314.7 1947
Physical Mechanism of [crystal] Rectifiers and Transistors—E. Spenke. (*Z. angew. Phys.*, vol. 5, pp. 472–480; Dec., 1953.) A survery of known types, based on electronic-semiconductor theory.

621.314.632:546.289:537.312.6 1948
Thermal Effects in Point-Contact Rectifiers—H. L. Armstrong. (*Jour. Appl. Phys.*, vol. 25, p. 136; Jan., 1954.) Correction to paper abstracted in 1242 of April.

621.314.7 1949
Irradiation of Transistors—C. D. Florida, F. R. Holt, and J. H. Stephen. (*Nature* (*London*), vol. 173, pp. 397–398; Feb. 27, 1954.) Measurements have been made on point-contact transistors with collector current limited by an external resistance. When the transistor was given a dose of about 10^{14} neutrons/cm^2, the decay of collector current after cessation of emitter current was speeded up, with little or no deterioration of the other parameters. The improvement appears to be permanent, and is attributed to the production of recombination centers such as lattice defects in the bulk material.

621.314.7 1950
Joining Solutions at the Pinch-Off Point in "Field-Effect" Transistor—R. C. Prim and W. Shockley. (*Trans. I.R.E.* PGED-4, pp. 1–14; Dec., 1953.) In a field-effect transistor with the drain connection biased beyond "pinch-off" in respect to the gate, the potential distribution is difficult to determine by analytic methods. An approximate solution is obtained which is valid over the whole length of the channel, and which completes one aspect of theory developed in 879 of 1953 (Shockley).

621.314.7 1951
Water Vapor and the "Channel" Effect in *n-p-n* Junction Transistors—R. H. Kingston. (*Phys. Rev.*, vol. 93, pp. 346–347; Jan. 15, 1954.) Experiments are reported which indicate that the conductance of the channels described by Brown (166 of January) depends on the water-vapor pressure. The mechanism of the effect is discussed.

621.314.7 1952
Behavior of Germanium-Junction Transistors at Elevated Temperatures and Power-Transistor Design—L. D. Armstrong and D. A. Jenny. (Proc. I.R.E., vol. 42, pp. 527–530; March, 1954.) Limitations on the operation of transistors at high temperatures, resulting from the increased thermal production of hole-electron pairs, are discussed in relation to *n-p-n* and *p-n-p* types for a dissipation of about 1 w. Methods of cooling are described [see also 1247 of April (Giacoletto)] whereby satisfactory α values are maintained at collector currents >100 ma.

621.314.7 1953
Calculation of the Surface Recombination Current in the Junction Transistor Prepared by Fusion—J. Laplume. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1107–1109; March 8, 1954.) The geometry of this type of transistor is such that volume recombination is negligible in comparison with recombination at the junction surfaces. An expression for the recombination current is derived on the assumption of a Laplacian concentration of injected charge carriers in the base.

621.314.7 1954
Evaluation of the Current Gain in the Fused-Junction Transistor—J. Laplume. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1300–1301; March 22, 1954.) The current gain α depends on the recombination current, a formula for which was derived previously (1953 above), and on the emitter current I_e. An approximate expression is derived for I_e based on the geometry of the junctions and on certain assumptions regarding the distribution of concentration of charge carriers. It is concluded that (*a*) α increases with collector radius, (*b*) for constant collector radius α exhibits a maximum when the emitter is slightly smaller than the collector, (*c*) when the emitter is larger than the collector α decreases rapidly. An optimum value of α is derived.

621.314.7:546.289 1955
A *P-N-P* Triode Alloy Junction Transistor for Radio-Frequency Amplification—C. W. Mueller and J. I. Pankove. (*RCA Rev.*, vol. 14, pp. 586–598; Dec., 1953; Proc. I.R.E., vol. 42, pp. 386–391; Feb., 1954.) A transistor having low resistance and low capacitance is obtained by using a thick wafer of low-resistivity Ge and reducing this thickness, by drilling, to about 0.0005 inch at the region where the junctions are formed. A gain of up to 39 db at 455 kc can be obtained using a neutralized circuit, and up to 12 db at 10 mc without neutralization. Characteristics are compared with those of the Type TA-153 transistor described by Law et al. (876 of 1953).

621.383 1956
Development of Photoemissive Receivers for the Far Ultraviolet Region. Photocells and Electron Multipliers—V. Schwetzoff. (*Rev. gén. Élect.*, vol. 63, pp. 71–96; Feb., 1954.) A detailed study of methods of preparation of photocathode and report of results for photocells and electron multipliers for operation at wavelengths below 2,000 Å. Prolonged heat treatment produced cathodes with very good characteristics. Ag-O-Cs and Sb-Cs photocathodes and Cu-Be electrodes for electron multipliers were specially studied. 199 references.

621.383.27:621.311.6 1957
Constancy of Photomultiplier Gain—R. Wilson. (*Jour. Sci. Instr.*, vol. 30, pp. 472–474; Dec., 1953.) An examination is made of features of the supply circuit which affect the gain of the photomultiplier, both focused and unfocused types being considered. Supply-circuit modifications are suggested by which the gain can be stabilized.

621.383.27:621.396.822 1958
Origin of Large-Amplitude Pulses in Photomultiplier Background Noise—Y. Koechlin, I. Pelchowitch and A. Rogozinski. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 660–662; Feb. 8, 1954.) Measurements on an E.M.I. Type-6260 photomultiplier at low temperature indicate that large-amplitude background pulses are to a great extent attributable to scintillations of the glass, caused by various radiations.

621.383.5 1959
Modification of the Spectral Response Curve of a Photocell due to Illumination Fatigue—G. Blet. (*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 578–579; Feb. 1, 1954.) Measurements made before and after a fatiguing illumination (1251 of April) indicate that the response is diminished throughout the spectrum without any particular manifestation in the neighborhood of the illumination wavelength, and that the degree of fatigue is independent of this wavelength; the effect is most marked at the long-wave end of the spectrum.

621.385.029.63/.64 1960
Some Remarks on Thermal Effect in an Electron Beam on the Gain of Traveling-Wave Tubes—K. Udagawa and M. Sumi. (*Jour. Appl. Phys.*, vol. 25, pp. 135–136; Jan., 1954.) Comment on 2580 of 1951 (Parzen and Goldstein).

621.385.029.63/.64 1961
A Wideband Power Mixer Tube—H. R. Johnson. (*Trans. I.R.E.*, PGED-4, pp. 15–27; Dec., 1953.) The mixer tube consists of a short input helix, a drift tube and an output helix arranged successively along an electron beam and surrounded by a magnetic field. The input helix and drift tube are maintained at low potential with respect to the cathode, the output helix is operated at higher potential. The valve thus combines low modulating-voltage requirements with high power output over a wide frequency band. Expressions are developed for carrier input power, sideband output power and modulating power. Results for experimental valves operating at about 3 kmc and having Pierce-type guns with L cathodes are reported.

621.385.029.63/.64 1962
High Power Traveling-Wave Tube Gain and Saturation Characteristics as a Function of Attenuator Configuration and Resistivity—J. J. Caldwell, Jr. (*Trans. I.R.E.*, PGED-4, pp. 28–32; Dec., 1953.) Preliminary results of tests on three tubes with different distributed attenuator configurations show that knowledge of loss alone is insufficient to determine attenuator performance and that the effect of loss on circuit impedance and phase velocity must also be known. For a given configuration, the attenuator with the higher resistivity gave the better performance.

621.385.029.63/.64:621.372.221 1963
Filter-Helix Traveling-Wave Tube: Part 1—The Filter Helix, New Circuit Element for Traveling-Wave Amplifiers and Oscillators—W. J. Dodds and R. W. Peter. (*RCA Rev.*, vol. 14, pp. 502–532; Dec., 1953.) A filter helix consists essentially of a helical transmission line with periodic inhomogeneities which cause reflections. It is basically a special type of a one-dimensional lattice. Various possible constructions are discussed, and three types are distinguished. See also 1524 of 1953 (Dodds et al.).

621.385.029.64/.65 1964
Wave Propagation along a Magnetically-Focused Cylindrical Electron Beam—W. W. Rigrod and J. A. Lewis. (*Bell Syst. Tech. Jour.*, vol. 33, pp. 399–416; March, 1954.) Theory given by Fletcher (1811 of 1950) for a beam in which the electrons move in straight lines (corresponding to infinitely strong magnetic field) is extended to the case where the electrons follow spiral paths (Brillouin flow, corresponding to finite magnetic field). The field equation is solved for the cases where the beam is surrounded (*a*) by a helix, and (*b*) by a drift tube. The gain constant for the cylindrical beam with Brillouin flow is greater than that of a similar beam with rectilinear flow but less than that of a thin hollow cylindrical beam with rectilinear flow, for the same radius, current, and voltage.

621.385.029-64/.65:513.647.1:621.317.336 1965
Note on Helix Impedance Measurements using an Electron Beam—D. A. Watkins and A. E. Siegman. (*Jour. Appl. Phys.*, vol. 25, p. 133; Jan., 1954.) Addendum to 3753 of 1953.

621.385.032.216 1966
A Hollow Beam Cathode—G. E. Mueller. (*Trans. I.R.E.*, PGED-4, pp. 33–36; Dec., 1953.) Tests on hollow cathodes of various volumes and apertures show that hollow electron beams with current densities of 25 A/cm² can be produced at a temperature of 1100 degrees C. The emission comes from the coating on the aperture edge, the large values of current being due to continual replenishment of the active Ba at this edge from the reservoir within the cavity. Cathode current is nearly proportional to anode potential. This departure from the 3/2 power law can be explained on the basis of the cathode geometry.

621.385.032.216 1967
A Study of the Evaporation Products of Alkaline-Earth Oxides—I. Pelchowitch. (*Philips Res. Rep.*, vol. 9, pp. 42–79; Feb., 1954.) Report of a comprehensive investigation of the evaporation characteristics of oxide-cathode materials; a mass-spectrometer method was used. BaO evaporates mainly in the form of the oxide when coated on Pt or Ni or coated in admixture with SrO and CaO on Pt. At high temperatures the BaO/Pt system gives off Ba_2O_2 ions. With the SrO/Pt and CaO/Pt systems the main evaporation product is the free element, accompanied by the oxide and the singly ionized metal ions. In the BaO/Ta system the main product is free Ba, accompanied by the oxide and by singly ionized Ba ions. The evaporation-rate/temperature curves exhibited critical points, whose existence was confirmed by resistance measurements on the oxides. The influence of the different base metals was demonstrated.

621.385.032.216 1968
The Properties of Oxide Cathodes regarded as Mixed Semiconductors—J. Ortusi. (*Ann. Radioélect.*, vol. 9, pp. 3–36; Jan., 1954.) After activation, the BaO layer is a mixed semiconductor, i.e. with both ionic and electronic conduction. The ionic conduction, which is due to the mobility of the oxygen ion, reacts on the electronic conduction, and modifies the emission characteristics. The electronic conduction is of *n* type, based on two groups of donors, the first consisting of Ba atoms, the second of F centers.

621.385.032.216:[546.42-31+546.431-31 1969
The Thermionic Emission from Thin Films of Barium and Strontium Oxide—J. Woods and D. A. Wright. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 74–76; Feb., 1954.) An experimental investigation is reported. The four principal results are: (*a*) emission from an evaporated film of BaO is a maximum at a layer thickness of 10^{-5} cm and, for an activated film, is approximately equal to that from a sprayed BaO cathode; (*b*) SrO behaves in a similar way, with a lower level of emission; (*c*) emission from BaO sprayed on SrO is similar to that from BaO alone; (*d*) SrO evaporated on BaO gives greater emission than either BaO or SrO alone, the dc emission for a SrO layer of thickness 10^{-5} cm being 9 mA/cm² at 550 degrees C.

621.385.1 1970
Valve Noise Produced by Electrode Movement—P. A. Handley and P. Welch. (PROC. I.R.E., vol. 42, pp. 565–573; March, 1954.) The effects of faulty electrode positioning and resonant vibrations of electrodes are distinguished. Expressions are derived for the resonance frequencies of the electrodes in terms of the constructional details; use of the results to improve valve design is discussed. Methods of measuring tube noise are indicated.

621.385.2/.3 1971
Self-Heating Thermionic Tubes—E. G. Hopkins. (*Proc. IEE*, part III, vol. 101, pp. 77–83; March, 1954.) Three experimental types of tube with mutually bombarding oxide cathodes (3209 of 1950) are described, namely (*a*) a diode in which the resistance is controlled by varying the spacing between the cathodes, the tube thus constituting a variable resistor, (*b*) a triode having a grid midway between the cathodes and functioning as a variable resistor or power oscillator, and (*c*) a diode in which the heated cathode area depends on the current, resulting in stabilization of the voltage between the cathodes. All three types operate direct from 240-v 50-cps mains.

621.385.2 1972
Transit-Time Effects in the Diode under Retarding-Field Conditions—F. W. Gundlach. (*Philips Res. Rep.*, vol. 8, pp. 419–426; Dec., 1953. In German.) Transit-time effects are taken into account in the calculation of the convection current at the anode. A calculation of the admittance shows the large contribution by electrons which do not reach the anode. Errors in the results of Knol and Diemer (290 of 1953) are noted.

621.385.2:537.525.92 1973
Approximate Solutions of the Space-Charge Problem for Some Unusual Electrode Geometries—H. F. Ivey. (*Jour. Appl. Phys.*, vol. 24, pp. 1466–1472; Dec., 1953.) The expressions for the space-charge-limited current for coaxial cylindrical or concentric spherical electrodes are usually given as functions of R/r. By applying the technique suggested by Matricon & Trouvé (1514 of 1951), expressions are obtained for R'/r, where R' is the radius of the equivalent cylindrical or spherical collector and r the radius of the emitter, for 27 different electrode arrangements. The results are tabulated.

621.385.3 1974
Development of a New Premium Twin Triode—H. E. Stumman and J. W. Ritcey. (*RCA Rev.*, vol. 14, pp. 482–491; Dec., 1953.) Description of the construction and manufacturing control of the Type 6101 tube, which is electrically and mechanically interchangeable with the Type 6J6 tube, but is specially designed for reliability.

621.385.832 1975
Viewing Storage Tube with Halftone Display—M. Knoll, P. Rudnick, and H. Hook. (*RCA Rev.*, vol. 14, pp. 492–501; Dec., 1953.) The construction and operation is described of a c.r. storage tube of the transmission-control type [3760 of 1953 (Smith and Brown)]. Writing speeds up to 3×10^6 dots/sec and an image persistence of $\sim$10 minutes for half-tone pictures have been obtained using a pulse restoring method.

621.385 1976
Electron Optics. [Book Review]—O. Klemperer. Publishers: Cambridge University Press, 471 pp., 50 s. 1953. (*Nature* (*London*), vol. 173, p. 182; Jan. 30, 1954.) An authoritative work, for those engaged in design and research in the field of electron optics as well as for physicists working in other fields.

how to stop an h-blast

WANTED: a camera to stop the action of a nuclear explosion at a pre-selected microsecond, with high quality image-definition . . . that was the problem handed by the AEC and its Los Alamos Scientific Laboratory to the Boston firm of Edgerton, Germeshausen & Grier, Inc. EG&G solved it by inventing the non-mechanical Rapatronic shutter . . . employing the Faraday Effect of magnetically rotating the plane of polarized light as it traverses an optical element . . . and relying on HELIPOT* precision potentiometers and DUODIAL* turns-counting dials for sensitivity setting and calibration.

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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 12A)

Power Oscillator

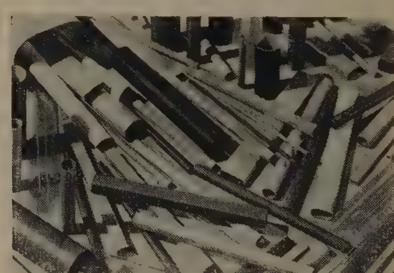
The Industrial Test Equipment Co., 55 E. 11 St., New York 3, N. Y., has introduced the Power Oscillator Model 1040. This unit provides a frequency of either 400 cps or 1,000 cps. Three watts of undistorted power (less than 1 per cent) are provided at various output impedance levels. A control on the front panel allows for a continuously variable output from 0

to 120 volts. Frequencies are factory-set to 0.25 per cent and are maintained with high stability even with line-voltage variations. The Model 1040 Oscillator is useful as a power source for all types of bridges and test set-ups. An isolated output transformer permits its use in modulation applications, and its high stability makes it a good frequency standard.

Power requirements are 105–125 volts, 60 cps. The unit weighs 10 pounds, and its dimensions are: height $5\frac{11}{16}$ inches, width 9 inches, depth $6\frac{1}{8}$ inches.

Coil Forms

Any coil form specifications in shape, size, length, I.D. or O.D., can now be met to within critical tolerances and supplied in quantity at low unit cost from **Precision Paper Tube Co.,** Dept. PIN, 2035 W. Charleston St., Chicago, Ill. Sizes from fractional inch to 9-inch inner diameter can be furnished without extra tooling charge.



According to the announcement, specially designed high production machinery which permits unusually close control during manufacture makes this possible.

To meet varying electrical and mechanical requirements, the forms may be wound from wide range dielectrical materials such as kraft, fish paper, acetate or combinations.

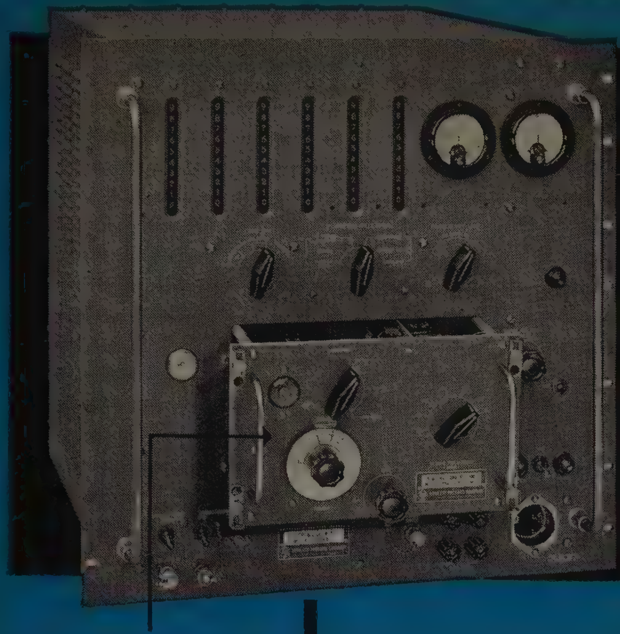
(Continued on page 80A)

Industrial Engineering Notes

(Continued from page 63A)

ment" which was held in San Francisco, Calif., April 19–20. The two-day conference was sponsored jointly by the Air Force Department and the Stanford Research Institute to permit scientists, businessmen, engineers, and military officials to exchange ideas on automation of electronic production. The keynote speaker was Air Force General Tom C. Rives, Manager of the General Electric Co. advanced electronics center at Cornell University. . . . **The Department of Defense recently announced that a radar mapping device known as the plan position indicator (P.P.I.) has been patented.** The patent was issued to Dr. Robert N. Page, electronics expert of the Naval Research Laboratory. The radar device was developed during the war and reportedly allows gunners and bombers to hit unseen targets, detect enemy ships and aircraft and aids in the navigation of Navy ships in bad weather. The plan position indicator is standard equipment on most commercial and military radar sets, it was reported. . . . **The Armed Forces Communications Association conference held in Washington May 6 and 7 saw the unveiling of the only instructional television system developed specifically for training purposes by the U. S. Navy.** The new ITV system, the result of five years of research, utilizes

(Continued on page 70A)



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Interval 1 μsec to 100 days
Period 0 cps to 10 kc

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-hp- 525B Frequency Converter

-hp- 526A Video Amplifier

-hp- 526B Time Interval Unit

-hp- 524B ELECTRONIC COUNTER

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Model 524B offers direct, instantaneous, automatic readings requiring no calculation, interpolation or complex instrument set-up. It has high sensitivity, high impedance, and its operation is so simple and dependable it can be used readily by non-technical personnel. Resolution is 0.1 μsec, and accuracy is 1/1,000,000 ± 1 count. Construction throughout is of highest quality components in a compact militarized design.

The new Counter with Plug-In Units gives you more range, more convenience, smaller size and lower cost than any commercial instrument combination ever offered. With this one compact equipment, you readily measure transmitter and crystal oscillator frequencies, time intervals, pulse lengths, repetition rates, frequency drift; make high accuracy ballistics time measurements or high resolution tachometry measurements, or use as a precision frequency standard giving convenience and flexibility not provided in the usual primary standard.

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BASIC COUNTER

The basic *-hp-* 524B Counter unit measures frequency from 10 cps to 10 mc with accuracy of ± 1 count ± stability, reading direct in kc; or measures period from 0 cps to 10 kc with accuracy of ± 0.3% reading direct in seconds, milliseconds or microseconds. Eight-place registration, short term stability 1/1,000,000, display time variable 0.1 to 10 seconds. $1,890.00

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-hp- 525A Frequency Converter extends Counter's range to 100 mc, maintains accuracy, and increases Counter's video sensitivity to 0.1 volts through basic 10 cps to 10 mc range. $225.00

-hp- 525B Frequency Converter like 525A but extends Counter's range from 100 to 220 mc at 0.25 volts sensitivity. $225.00

-hp- 526A Video Amplifier increases Counter sensitivity between 10 cps and 10 mc to 10 millivolts for low level frequency measurement. $125.00

-hp- 526B Time Interval Unit measures interval 1.0 μsec to 100 days with accuracy of 0.1 μsec ± 0.001%, reading direct in seconds, milliseconds or microseconds. Start, stop triggering in common or separate channels, through positive or negative going waves. $150.00 (Plug-in units supplied in aluminum storage case).

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Industrial Engineering Notes

(Continued from page 68A)

simplified equipment specially designed for use by ordinary classroom and laboratory instructors, heretofore made too expensive by costly studio equipment and personnel, the Navy announced. The prototype model of the ITV system on display featured one orthicon camera with provision for including two more cameras. The camera, with its complete receiving and transmitting equipment and sound system has been housed in a small, desk-like console easily moved from one classroom to another. From this console, the Navy said, the TV program can be transmitted by cable to as many as 100 different TV receivers, located in as many different areas. A Department of Defense announcement stated that the new ITV system can be used for presenting demonstrations or illustrated lectures by master teachers or specialists to as many as 100 classes at the same time. And for technical skill training, it would be possible for each student to have a TV receiver at his work bench and follow the instructors step-by-step.

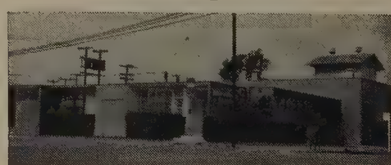
Federal Personnel

Capt. W. H. G. Finch (USNR) has been transferred from the Bureau of Ships to the post of Assistant Chief of the Office of Naval Research. He also has been designated Patent Counsel for the Navy. Capt. Finch previously was Special Assistant for Counter-measures and Patent Matters to the Assistant Chief of the Bureau of Ships for Electronics.

Standardization

The Department of Defense issued a notice recently pointing out that a revision of the military standard on electronic and electrical symbols (MIL-STD-15A) now is in effect. Five basic changes reportedly have been made in the new standard compared with the previous one which was dated October 1948. It was pointed out that as the result of close co-operation with industry there are no conflicts with American Standard Y32.2, Graphic Symbols for Electrical Diagrams, which is expected to be issued shortly. MIL-STD-15A is available through the Superintendent of Documents, Government Printing Office, Washington 25, D. C.

Plant Expansion



T. R. Fowler, President of **Mesa Plastics Co.,** 11751 Mississippi Ave., Los Angeles 35, Calif., announces the opening of a new and enlarged plant. The new installation will include all present offices and production facilities, plus several separate facilities for the production of a new and improved line of Diallyl Phthalate molding compounds.

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Professional Group on Electron Devices

Developments in the field of electron devices have probably had a greater influence on radio-electronic progress than the developments of any other field. For it is the electron device which is the active, working element around which the rest of the equipment is built. It is this component which generates, modulates, converts, detects, and amplifies signals, and which enables the apparatus to perform its prescribed functions.

It is not surprising, therefore, that a great deal of effort has been directed toward improvements in electron tubes and semiconductors, and that the resulting advances have received close and widespread attention. The rewards have been impressive. Work on cathode-ray tubes, for example, was a prime factor in the successful evolution of television. The advent of klystrons, magnetrons, and traveling-wave tubes made possible the development of equipment capable of operating in the microwave portion of the spectrum. The transistor offers new possibilities which are only now beginning to be realized.

IRE members who wished to keep abreast of the many developments in this important and rapidly moving field were greatly aided by the formation of the IRE Professional Group on Electron Devices in 1951. The Group immediately began to organize technical sessions in its field at many of the national meetings held during the year. This was followed by the formation of Group Chapters throughout the country which collaborate with IRE Sections to sponsor meetings locally.

Perhaps of greatest importance, the Group began publishing its own technical publication, called *Transactions,* which is distributed free to all members who pay the assessment fee of $2.00. The *Transactions* is now issued quarterly to some 1300 members, and has become an invaluable source of information on the latest technical developments in the field of electron devices.

W. R. G. Baker

Chairman, Professional Groups Committee

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PROFESSIONAL GROUP MEETINGS

Aeronautical and Navigational Electronics

The Los Angeles Chapter of the Professional Group on Aeronautical and Navigational Electronics met on April 13 at the Institute of Aeronautical Sciences Building in Los Angeles. Henry Blanchard of Stanford Research Institute spoke on "Effects of Helicopter Rotor Modulation on VOR Receiver," and Robert C. Twomey of Douglas Aircraft Company spoke on "Drag and Weight Considerations in Airborne Antenna Design." The meeting was co-sponsored by the Professional Group on Antennas and Propagation.

The New York Chapter of the Group met on March 11 at the General Electric Company Auditorium under the chairmanship of R. J. Bibbero. Dr. K. C. Black, Chairman of the PGANE, was the speaker. A discussion of Chapter organization and aims and appointment of an ad hoc committee for organization followed the speech.

Audio

The Boston Chapter of the Professional Group on Audio met on April 8, 1954 at Massachusetts Institute of Technology under the chairmanship of J. A. Kessler. Derwent M. A. Mercer of the Department of Physics, University of Southampton, England, spoke on "Organ Pipe Tones." A number of members of the American Guild of Organists were guests at the meeting.

The Cleveland Chapter of the Group met on April 29 at Station WHK in Cleveland, under the chairmanship of Herber H. Heller. Norman C. Pickering, President of Pickering and Co., presented a paper entitled "Important Details in Sound Reproduction."

The Philadelphia Chapter met on April 22 at the Franklin Institute in Philadelphia under the chairmanship of M. S. Corrington. Dr. Winston E. Kock, Director of Acoustics Research, Bell Telephone Laboratories, spoke on "Polarized Airborne Sound Waves." Dr. Kock was assisted by F. H. Harvey in the demonstrations.

The Phoenix Chapter met on March 5 under the chairmanship of Allen M. Creighton to hear W. E. Kock of Bell Telephone Laboratories speak on "Physics of Music and Hearing." The following officers were elected at this meeting for the coming year: Chairman, North C. Ham, and Secretary-Treasurer, Andrew B. Jacobsen.

Electron Devices

The New York and Long Island Chapter of the Professional Group on Electron Devices met on February 11 at the General Electric Auditorium in New York under the chairmanship of C. E. Fay. Dr. R. G. E. Hutter, Branch Head of Sylvania Electric Products Inc., presented a paper entitled "The Relationships Among Trav-

(Continued on page 78A)

PROFESSIONAL GROUP MEETINGS

(Continued from page 74A)

eling-Wave Tubes." On November 24, 1953 the Group met to hear Dr. Wen Yuan Pan of RCA speak on "Investigation of UHF Television Amplifier Operation"; and Donald E. Nelson of RCA on "A High-Power CW Magnetron."

The Philadelphia Chapter of the PGED met on March 8 at the University of Pennsylvania under the chairmanship of Dr. E. I. Hawthorne, to hear Edward O. Johnson, Research Engineer at RCA Laboratories, speak on "A New Grid-Controlled Thyratron." The paper was discussed by the audience.

Electronic Computers

The Philadelphia Chapter of the Professional Group on Electronic Computers met on April 20 at the Franklin Institute Auditorium, under the chairmanship of John M. Broomall. J. H. Dever, Minneapolis-Honeywell Regulator Co., presented a paper entitled "Computing the Behavior of a Nuclear Reactor." Mr. Dever had on display a working model of this computer with recorders to show reactor power level, control period, and control rod position.

The San Francisco Chapter of the Group met on April 7 at Corey Hall, University of California at Berkeley, under the chairmanship of T. H. Meisling. William H. Kautz of Stanford Research Institute presented a paper on "Data Encoding for Digital Computers," and Abraham Katz of Marchant Research Inc. presented a paper entitled on "Digital Control Systems."

The Washington, D. C. Chapter met on May 5 at the PEPCO Auditorium in Washington. The presiding officer was Ralph J. Slutz, Secretary. Mr. Jack Rabinow, Consulting Engineer, presented a paper entitled "Reading Machines."

Engineering Management

The Philadelphia Chapter of the Professional Group on Engineering Management met on January 14 at the Franklin Institute in Philadelphia, under the chairmanship of A. D. Emurian. Dr. Melville Hopkins, Professor at Pennsylvania State University, spoke on "Human Relations in Engineering Management." Dr. Hopkins' subject was intensely interesting to this particular group.

On April 20 the Philadelphia Chapter also met to hear a paper presented by John J. Slattery, Director of Counter Measures at Evans Signal Laboratory. The paper was titled "The Function and Management Problems of the Government Laboratory with Particular Reference to the Evans Signal Laboratory."

Information Theory

The Los Angeles Chapter of the Professional Group on Information Theory met on April 8 at the Institute for Numerical Analysis, University of California, under the chairmanship of Robert B. Bennett. The speakers for the evening were

(Continued on page 80A)

PROFESSIONAL GROUP MEETINGS

(Continued from page 78A)

Harold Davis of UCLA on "Relations Between Alternate Definitions of Information Content," and Robert M. Steward, Jet Propulsion Laboratories, on "Some Problems in System Analysis."

MEDICAL ELECTRONICS

The San Francisco Chapter of the Professional Group on Medical Electronics met on March 4 under the chairmanship of Albert J. Morris. Hans Hollman, Research Scientist, USN Test Center, Pt. Mugu, presented a paper entitled "Electrocardiography."

RADIO TELEMETRY AND REMOTE CONTROL

The Los Angeles Chapter of the Professional Group on Radio Telemetry and Remote Control met on April 20 at the IAS Building in Los Angeles, under the chairmanship of John R. Kauke. M. W. Brainard and A. T. Puder spoke on "Airborne Primary Power for Telemetering and Rotating Power Supplies." R. S. Gardner, and E. F. Errico spoke on "Silver-Zinc Alkaline Batteries and Mercury Batteries." Charles Burrell and Dr. S. Herbert spoke on "Dry Batteries for Special Applications."

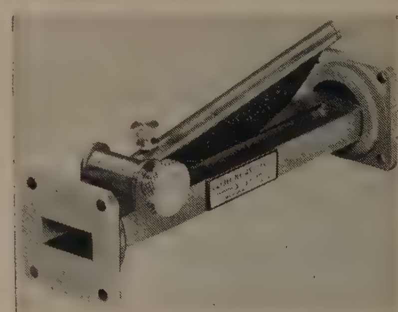
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 66A)

X-Band Flap Attenuator

A new, economically-priced flap attenuator manufactured by **Guideline Associates,** 47 State St., Newark, N. J., provides a method of adjusting power levels and isolating oscillators from the pulling effects of variable loads. With a



fully inserted flap it may be employed as a load. The unit operates over the X-Band spectrum from 8,200 mc to 12,400 mc. For attenuation values less than 20 db maximum vswr is less than 1.15.

A simple screw mechanism positively varies the attenuation continuously from 0 to over 30 db. An auxiliary lock device, when used, offers insurance against variations in attenuation due to vibration, shock and curious "knob-turners."

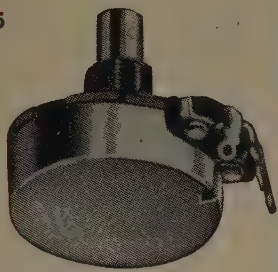
(Continued on page 111A)

RESISTORS...FOR EVERY NEED

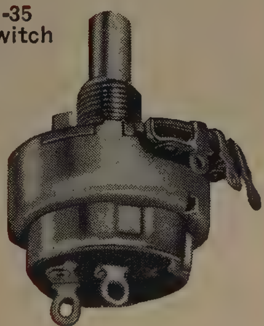
Meets JAN-R-94
type RV3

1/2 watt 1-1/8″ diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94.

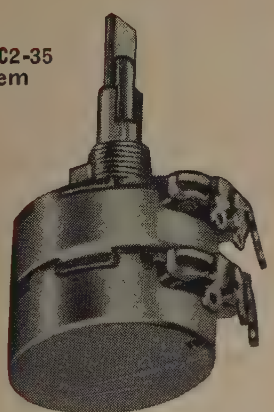
TYPE 35



TYPE GC-35
With Switch



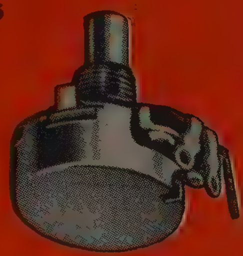
TYPE C2-35
Tandem



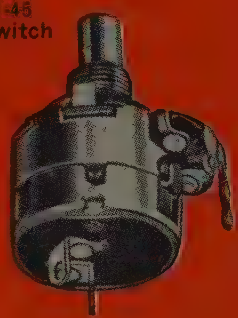
Meets JAN-R-[illegible]
type [illegible]

1/4 watt 15/16″ diameter variable composition resistor. Also available with other special military features not covered by JAN-R-94.

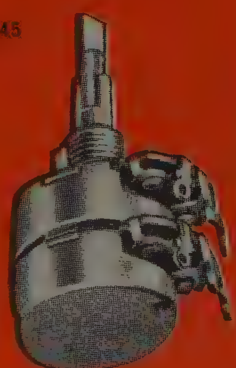
TYPE 45



TYPE GC-45
With Switch



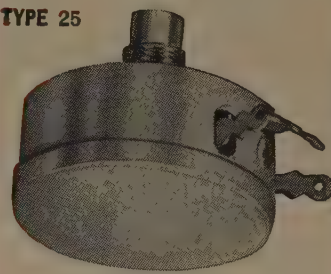
TYPE C2-45
Tandem



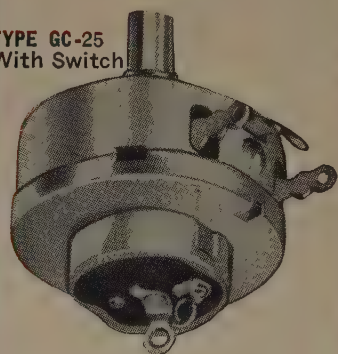
Meets JAN-R-19
types RA25 and RA30

4 watt 1-17/32″ diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19.

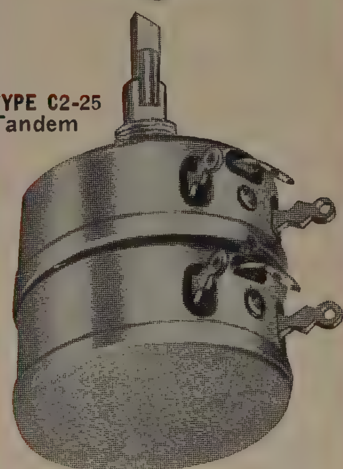
TYPE 25



TYPE GC-25
With Switch



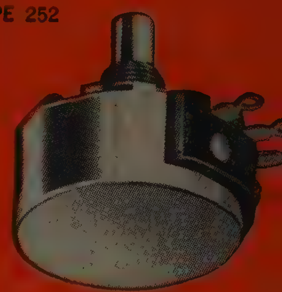
TYPE C2-25
Tandem



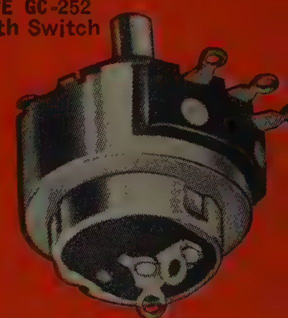
Meets JAN-R-[illegible]
type [illegible]

2 watt 1-17/64″ diameter variable wirewound resistor. Also available with other special military features not covered by JAN-R-19.

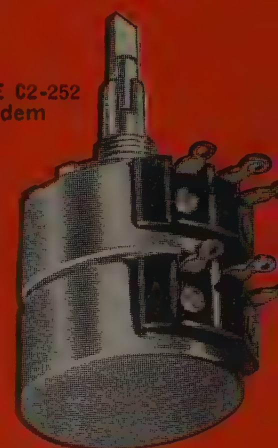
TYPE 252

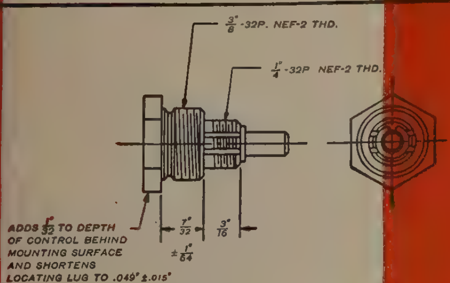


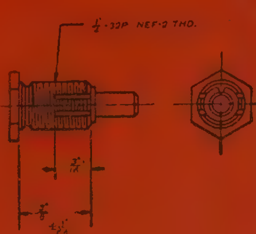
TYPE GC-252
With Switch



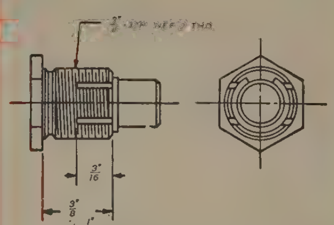
TYPE C2-252
Tandem

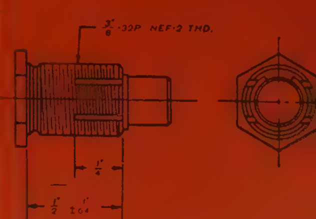






LOCKING BUSHINGS FOR TYPE 65 CONTROL.





LOCKING BUSHINGS FOR CONTROL TYPES 25, 252, 95, 35, 90, 45.

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Contributors

For a photograph and biography of R. H. Baker, whose paper appears on page 1152 of this issue, see page 1191 of the September, 1953 issue of the PROCEEDINGS OF THE I.R.E.

L. E. Barton (M'29–SM'43) is a senior member of the Research Staff of the RCA Laboratories in Princeton. His paper appears on page 1062 of this issue.

L. E. BARTON

He was born in Fayetteville, Ark., November 7, 1897, and served in the armed forces during World War I. He received the B.E.E. degree in 1921 and the E.E. degree in 1925 from the University of Arkansas.

Mr. Barton served as an instructor in Mechanical Engineering at the University of Arkansas from 1921 to 1925 and returned as an Associate Professor in 1927 for two years after a two year student training course at the General Electric Co. He was employed by RCA in Camden, N. J. as a radio engineer in 1929. Except for the period from 1936 to 1939 with Philco he was with RCA Victor at Camden in various capacities as a radio engineer and engineering supervisor until 1946.

Approximately 40 patents on radio devices have been issued to Mr. Barton, some of which pertain to class B audio amplifiers. The class B audio amplifier was first used for high level broadcasting station modulation at the University of Arkansas late in 1928. This type of modulation is now used in practically all medium frequency broadcasting stations. Other patents pertain to Sonar, Radar, air navigation systems, color television, and transistors. Several articles pertaining to class B audio amplifiers were written in the early 30's for radio publications including QST and the PROCEEDINGS OF THE I.R.E.

Mr. Barton received letters of appreciation from the Navy for work done on Sonar and for work done at the Naval Research Laboratory in Washington during the second World War. He received a citation for distinguished alumnus from the University of Arkansas in 1947.

Mr. Barton is a member of Tau Beta Pi and Sigma Xi.

(Continued on page 86A)

New Laboratory

Techniques, Inc., Bldg. 20, Santa Barbara Airport, Goleta, Calif., a new research and development laboratory, has been established to provide facilities for design of radar, computers, and servomechanisms. The laboratory will also handle antenna or similar outdoor work and problems associated with aircraft, such as installation and airborne testing. The principal engineers leading the organization are Carl A. Wiley and Ben R. Gardner.

Contributors

(Continued from page 84A)

E. A. Blasi (A'36–SM'46) whose paper appears on page 1179 of this issue, recently accepted a position as a member of the technical staff, Research and Development Laboratories, of the Hughes Aircraft Company.

E. A. Blasi

He was born in New Rochelle, N. Y., in 1913. He received the B.S. degree in electrical engineering from New York University in 1934.

From 1934 to 1940 he was engaged in radio receiver design with the Radio Receptor Company and the General Electric Company. From 1940 to 1942 he was a test engineer in power distribution networks with the Westchester Lighting Company. He joined the Radar Laboratory, at Wright Field, in 1942, and was commissioned in the United States Army in the fall of the same year. During his stay in the Army, he was in charge of the research and development of microwave test equipment for airborne radar bombing, gun-laying, and search systems. In 1946 he was appointed chief of the specialized microwave components unit of the Aircraft Radiation Laboratory at Wright Field, and a year later was appointed chief of the Antenna Design Section.

❖

For a photograph and biography of G. A. Boggs, whose paper appears on page 1144 of this issue, see page 761 of the April, 1954 issue of the Proceedings of the I.R.E.

❖

A. B. Bronwell (A'39–SM'43) whose paper appears on page 1117 of this issue is Professor of Electrical Engineering at Northwestern University and also holds the position of Executive Secretary of the American Society for Engineering Education and for five years was Editor of the *Journal of Engineering Education*. He is directing a project on microwave detection under a grant from the National Science Foundation.

A. B. Bronwell

He was born in Chicago, Ill. on August 18, 1909. He graduated from Illinois Institute of Technology with the B.S. and M.S. degrees in E.E. He also received an M.B.A. degree from the Northwestern University in 1947.

During the war he organized and di-

(Continued on page 89A)

Contributors

(Continued from page 86A)

rected an Army Signal Corps School at Northwestern University for advanced training in radio and communications. He also supervised a radar research project dealing with the development of microwave oscillators for Airborne radar units. In addition, he served as a consultant to Galvin Mfg. Co. and as special project engineer for Bell Telephone Labs.

Professor Bronwell has served on the RCA Fellowship Board, has been Chairman of the Education Committee of the IRE, Chairman of the Chicago Section of the IRE, and was one of the founders of the National Electronics Conference and served as its President in 1947. He is the author of a recent book, "Advanced Mathematics in Physics and Engineering," and is the co-author of "Theory and Application of Microwaves." In 1951, he was a member of the United States Mission on Engineering Education which visited Japan. He holds a patent on an all-electronic method of reproducing color television pictures.

Professor Bronwell holds membership in Tau Beta Pi, Eta Kappa Nu, Sigma Xi, AIEE, ASEE, and the American Economic Association.

❖

H. A. Chinn (A'42–SM'45–F'45) whose paper appears on page 1067 of this issue is currently Chief Engineer of the Audio-Video Division of the General Engineering Dept. of Columbia Television. He joined the company in 1932.

H. A. Chinn

He was born on January 5, 1906 in New York, N. Y. He received the S.B. and S.M. degrees in 1927 and 1929, respectively, from Massachusetts Institute of Technology where, from 1927 to 1932, he was a research associate.

For his work with the Office of Scientific Research & Development during the war years, he received a Presidential Certificate of Merit, the second highest award available to civilians. In 1950, he received the John H. Potts Memorial Award for outstanding achievement in the field of audio engineering.

Mr. Chinn is a Fellow in the Audio Engineering Society, the Institute of Radio Engineers, the Society of Motion Picture and Television Engineers and the Acoustical Society of America. He is a Member of the American Institute of Electrical Engineers, a licensed professional engineer and the author of "Television Engineering," one of the volumes in the McGraw-Hill television series.

(Continued on page 90A)

Contributors

(*Continued from page 89A*)

P. E. Fish (A'41–SM'46) joined in 1945 the Columbia Broadcasting System, Inc., where he is a senior engineer in the General Engineering Department of CBS Television engaged in the development, design and construction of audio and video broadcasting facilities. His paper appears on page 1067 of this issue.

P. E. Fish

He was born on Sept. 18, 1911 in Fort Worth, Texas. He received the B.A. degree in mathematics from William Penn College in 1935. From 1938 to 1942 he was employed by the United Broadcasting Company in Cleveland, Ohio, and in 1942 joined the scientific staff of the U. S. Navy Underwater Sound Laboratory at New London, Connecticut, where he was engaged in the development of submarine underwater sound devices.

Mr. Fish is a governor of the Audio Engineering Society.

❖

S. F. George (A'45–SM'51) whose paper appears on page 1159 of this issue has served as consultant and head of the mathematics staff in Radio Division III of the Naval Research Laboratory since 1948.

S. F. George

He was born in Mason City, Iowa, on July 21, 1920. He received his B.A. degree from the University of Iowa in 1942, where he majored in mathematics.

After serving for a short time in the actuarial department of the Lincoln National Life Insurance Company in Fort Wayne, Ind., Mr. George became associated with the U. S. Naval Research Laboratory in Washington, D. C. in 1943. His chief fields of interest are information theory, correlation theory, and noise.

Mr. George is a member of Phi Beta Kappa and the Scientific Research Society of America.

(*Continued on page 92A*)

Contributors

(Continued from page 90A)

R. E. Greenquist (S'48–A'49) joined the Radio Research Division of the Western Union Telegraph Company in 1948 and engaged in the tropospheric propagation experimental program then being conducted. Subsequently he worked on microwave radio relay systems development and the design of monitoring equipment for such systems. At the present time Mr. Greenquist is the project engineer responsible for research on antennas and propagation. His paper appears on page 1173 of this issue.

R. E. Greenquist

He was born in Chelsea, Massachusetts on March 30, 1921. He attended Northeastern University and was employed by the Allis-Chalmers Manufacturing Company as a student engineer until he entered the U. S. Air Force in 1942. Returning from military service in 1946, he transferred to Cornell University and in 1948 received a B.E.E. degree with a communications major.

❖

Walter Hitschfeld whose paper appears on page 1165 of this issue is an assistant professor of physics at McGill University, Toronto, Canada.

W. Hitschfeld

He was born in Vienna, Austria, in 1922. Since coming to live in Canada in 1941 he has graduated from the University of Toronto with a B.S. degree in engineering in 1946. He obtained the Ph.D. degree in physics at McGill University in 1950.

Since 1948 Dr. Hitschfeld has been associated with the Stormy Weather Research Group at the Defense Research Board in Ottawa and then at McGill University in Montreal. While with the Defense Research Board he has worked on problems relating to the mechanism of precipitation and the application of radar to weather studies.

Dr. Hitschfeld is a Fellow of the Royal Meteorological Society and an officer of its Montreal Centre. He is a member of the Canadian Association of Physicists and of Sigma Xi.

❖

J. D. Lawson whose paper appears on page 1147 of this issue joined the Atomic Energy Research Establishment in 1947.

(Continued on page 94A)

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Cylindrical assembly with 25 rings. Three wide rings accommodate large contact area brushes for high current capacity. Length 14″, O.D. approx. 5³/₈″.

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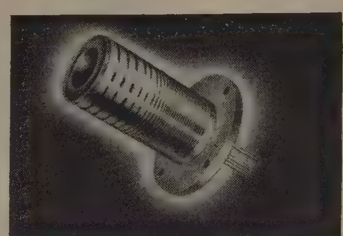
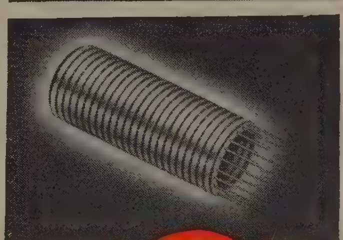


An assembly with 30 rings of various widths to accommodate various current requirements. Unit is approx. 4-5/16″ long, designed for flange mounting.



Cylinder type assembly approx. 3³/₄″ long with 24 hard silver rings. 1⁵/₈″ O.D. with wall thickness less than ¹/₄″.

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Contributors

(Continued from page 92A)

J. D. Lawson

After two years work on the design and performance of synchrotron accelerators, he turned to a general study of the radiation characteristics of high energy electron-accelerator targets, and methods of measuring the power flux in the X-ray beam. Apart from some work on the scattering of high energy nucleons, his main interest during the last three years has been in high power microwave klystrons.

He was born at Coventry, England, on April 4, 1923. In 1943 he graduated in Mechanical Sciences from Cambridge University.

After graduation he joined the Telecommunications Research Establishment at Malvern. Here he worked in the microwave antenna group until 1947.

❖

For a photograph and biography of I. L. Lebow, whose paper appears on page 1152 of this issue, see page 1191 of the September, 1953 issue of the Proceedings of the I.R.E.

❖

J. S. Marshall joined the physics department of McGill University at the end of World War II. He formed the Stormy Weather Research Group there, for studies in radar weather and cloud physics, and has this year been appointed a professor. His paper appears on page 1165 of this issue.

J. S. Marshall

He was born at Welland, Ontario, in 1911. He obtained a B.A. with honours in mathematics and physics from Queen's University in Kingston in 1931, followed by an M.A. from Queen's (1933) and a Ph.D. from Cambridge University in 1941, his research being in nuclear physics.

On the staff of the National Research Council of Canada from 1939 until 1945, Dr. Marshall did work in ballistics and operational research, the latter including project "Stormy Weather," a study of radar weather echoes in the Canadian Army Operational Research Group.

He is a Fellow of the Royal Society of Canada, a past president of the Canadian Association of Physicists, and chairman of Canadian Commission 2 of URSI.

(Continued on page 96A)

Contributors

(Continued from page 94A)

G. L. Matthaei (S'49–A'52) has been an instructor in the Division of Electrical Engineering of the University of California at Berkeley since 1951. His paper appears on page 1126 of this issue.

G. L. MATTHAEI

He was born in Tacoma, Wash., on August 28, 1923. He entered the University of Washington in 1941, but left during the period 1943 to 1946 for Army service in the U. S. and the Pacific Theater. He returned to Washington and received his B.S. degree in electrical engineering in 1948. He then received his M.S. at Stanford University in 1949, the degree of Engineer in 1951, and the Ph.D. degree in 1952.

He is a member of Tau Beta Pi and Sigma Xi.

❖

J. C. May (S'50–A'52), whose paper appears on page 1117 of this issue, was employed by the Cook Research Laboratories, Skokie, Ill., in their radar section in 1952. At present he is Technical Director of a radar field test program.

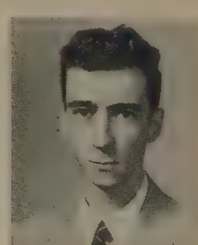
J. C. MAY

He was born September 24, 1926 in Manitowoc, Wis. He served in the U. S. Navy as an electronics technician from 1944 to 1946. In 1947 he entered Northwestern University where he received the B.S. degree in 1951 and the M.S. degree in 1952. While attending college he was employed by the Allen B. DuMont Laboratories under a co-operative training program.

Mr. May is a member of Eta Kappa Nu and Tau Beta Pi.

❖

R. E. McMahon whose paper appears on page 1152 of this issue is a member of the staff of Massachusetts Institute of Technology Lincoln Laboratory where he is currently engaged in research on transistor applications to digital computors.



R. E. McMAHON

He was born on July 1, 1928 in Lynn, Massachusetts. He received the B.S. degree in Electrical Engineering from Georgia Institute of Technology in 1952.

Mr. McMahon holds membership in Eta Kappa Nu and Tau Beta Pi.

(Continued on page 98A)

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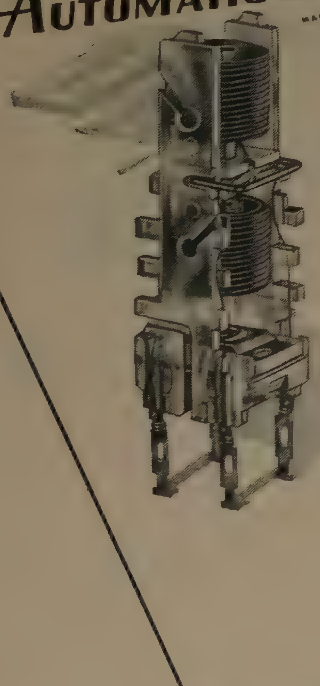

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Martin H Benedek

President

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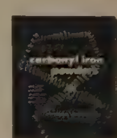

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Contributors

(Continued from page 96A)

R. B. Monroe (A'42–SM'46) joined the Columbia Broadcasting System, Inc. in 1934 and is currently a senior project engineer in the General Engineering Department of Columbia Television specializing in the development and design of audio and video systems for broadcasting application. His paper appears on page 1067 of this issue. During the war years, Mr. Monroe was associated with the Radio Research Laboratory, Harvard University, sponsored by the Office of Scientific Research & Development.

R. B. Monroe

He was born on October 17, 1908 in Brooklyn, N. Y. and attended Pratt Institute from 1937–42.

He served as head of the Planning Department, head of the Standards Laboratory, and assistant to the Executive Engineer. Mr. Monroe is co-author with George E. Sterling of the book "The Radio Manual."

❖

For a photograph and biography of G. F. Montgomery, whose paper appears on page 1184 of this issue, see page 488 of the February, 1954 issue of the Proceedings of the I.R.E.

❖

I. C. Nitz, Jr. whose paper appears on page 1117 of this issue is associated with Sylvania Electric Products Co., Microwave Tube Laboratory, Mountain View, Calif.

I. C. Nitz, Jr.

He was born in Evanston, Ill., on March 18, 1923. He entered the Technological Institute at Northwestern University in 1940, was employed for four quarters by the Public Service Company of Northern Illinois under the co-operative work program, and was awarded a B.S. degree by the University Senate in February 1944, when commissioned by the Navy. He was a commanding officer of an LSM before returning to Northwestern where he completed his B.S. degree in electrical engineering in December 1946.

Mr. Nitz worked with Childs and Smith, an architectural firm, as an electrical engineer for two years, and then was employed by Marshall Field and Company as a project engineer for nearly three years. He re-entered Northwestern in October, 1951 and received his M.S. degree in June 1953.

Mr. Nitz is a member of Tau Beta Pi and Pi Mu Epsilon.

(Continued on page 100A)

Contributors

(*Continued from page 98A*)

R. S. O'Brien (S'39–A'43–M'45) is on the staff of the Columbia Broadcasting System where he is now a senior project engineer in the General Engineering Department of the Columbia Television Division, specializing in the development, design, and construction of television video systems. His paper appears on page 1067 of this issue.

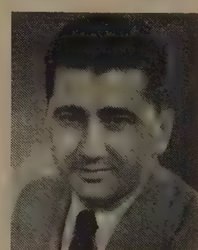
R. S. O'Brien

He was born on July 19, 1917 in Galt, Iowa. He received a B.S. degree in electrical engineering from the University of California in 1939 and did graduate study in electronics and television at Stanford University in 1939 and 1940. From 1941 to 1943 he did electronic instrumentation for Standard of California, Inc. From 1943 to 1945 he was at the Radio Research Laboratory, Harvard University, becoming a division leader.

He is a member of the Society of Motion Picture and Television Engineers, Eta Kappa Nu and Sigma Xi.

❖

A. J. Orlando joined the Radio Research Division of the Western Union Telegraph Company in 1951 where he engaged primarily in microwave antenna studies and experiments. He is presently engaged in microwave propagation experiments. His paper appears on page 1173 of this issue.



A. J. Orlando

He was born in Jersey City, New Jersey on August 5, 1924. In 1947 he attended Champlain College and transferred to Lehigh University the following year. In June 1951 he received the B.S. degree in electrical engineering from Lehigh University.

Mr. Orlando is continuing his studies in the Graduate School of Polytechnic Institute of Brooklyn.

He is a member of Eta Kappa Nu.

(*Continued on page 102A*)

For complete information about joining IRE Professional Groups see page 72A of this issue.

Contributors

(Continued from page 100A)

N. M. Schaeffer whose paper appears on page 1115 of this issue is at present a member of the nuclear engineering staff of Consolidated Vultee Aircraft Corporation, Fort Worth division.

N. M. SCHAEFFER

He was born on November 1, 1927, in Camden, Ark. He received the B.S. degree, and the M.S. degree in physics, in 1947 and 1949, respectively, from Louisiana State University. He then enrolled in the Graduate School of the University of Texas, where he received the Ph.D. degree in 1953. From 1950 to 1953 he was employed as a research physicist with the Defense Research Laboratory, University of Texas.

He is a member of the American Physical Society.

❖

H. F. Schulte (S'47–A'49) whose paper appears on page 1104 of this issue became project engineer in upper atmosphere research for the Engineering Research Institute of the University of Michigan in 1953.

H. F. SCHULTE

He was born on July 23, 1923 in Detroit, Mich. He attended the University of Michigan in 1941. He left for military service during the period from 1943 to 1946 at the Naval Research Laboratory. He was an engineering aide in the Electronic Special Research Division and he was engaged in radar beacon research and development. He returned to the University of Michigan following the war and received the BSE degree in electrical engineering in 1949.

Since he received his B.S. degree he has continued his graduate studies and has been engaged as a research and development engineer for the ERI of the University.

(Continued on page 104A)

MICROWAVE DEVELOPMENTS

Wheeler Laboratories is an engineering organization offering consulting, research, and development services in the fields of radio and radar.

The high-power rotary joint pictured above is a design based on principles originated in our laboratory. It is now being adapted for various waveguide sizes as a general-purpose component. It is designed to equal the pulse-power capacity of the largest permissible coaxial line and to hold the reflection within 1 db SWR over a frequency bandwidth of 20 percent.

At present, Wheeler Laboratories comprises a staff of twenty engineers under the personal direction of Harold A. Wheeler, with supporting facilities including a group of designers and a model shop.

A brief summary of our work will be sent on request, and comprehensive engineering reports on some of our developments are available. Inquiries are welcomed regarding your particular problems in microwave design and development.

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2SS1	2400–3650mc/s	2SX2	6250–7425mc/s
2SS1a	2400–4340mc/s	2SX1a	9500–11,250mc/s
2SC1b	4240–4910mc/s	2SX1a	8500–11,250mc/s
2SC1a	4240–5900mc/s	2SX1	8500–9660mc/s
2SC1	5100–5900mc/s	2SK1	15,300–17,700mc/s
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Contributors

(Continued from page 102A)

S. Sherr (SM'53) whose paper appears on page 1169 of this issue joined the General Precision Laboratory Incorporated in 1951 where he now holds the position of Associate Section Head of the Electronic Circuits Section, with specific responsibility for advanced development and special projects, including transistor, miniaturization and automation.

S. Sherr

He was born in New York City, N. Y. on March 23, 1918. He was graduated from the Juilliard School of Music in 1939 with the B.S. degree in Music Education, and from New York University in 1947 with the B.A. degree in Electronic Physics. He has done graduate work at New York University, Polytechnic Institute of Brooklyn, Wayne University, and Teachers College, Columbia University.

Mr. Sherr was employed by Western Electric Co. from 1942 to 1947 as a test equipment design engineer, and then as a circuit development engineer by Bendix Aviation Research Laboratories in 1948. During the period from 1948 to 1951 he was employed by Federal Telecommunications Laboratories as a senior development engineer.

Mr. Sherr is a member of Sigma Pi Sigma.

❖

H. S. Sicinski whose paper appears on page 1104 of this issue has been engaged as a research physicist with the ERI of the University of Michigan for the past four years. In this capacity he has assumed responsibility for the development and analysis of many aerodynamic studies associated with high altitude supersonic missiles and upper air physics.

H. S. Sicinski

He was born on March 30, 1924 in Chicago, Ill. After attending the University of Michigan for a year, he served three years in the U. S. Army. During this period, he attended the University of Wyoming under an A.S.T.P. and served with an NDRC unit in Florida developing automatic sampling equipment for micro-meteorological studies of mustard gas and other vesicants.

Following this he was employed by the Communications Equipment Corp., of Chicago. He returned to the University of Michigan in 1949 and received the BSE degree in physics and later the MS degree in physics.

Mr. Sicinski is a member of the American Institute of Physics.

(Continued on page 106A)

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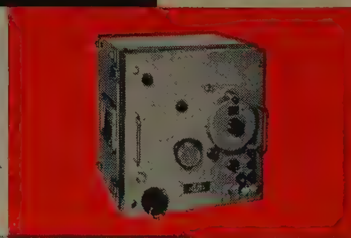
SIGNAL GENERATORS



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The Extremely High Frequency Polarad Signal Generator, for example, furnishes monitored power output as well as measures external signal strength and frequency.

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Frequency Range	Model Number	Output Power	Model Number	Power Output (Average)	Model Number	Sensitivity (Signal=Noise)	Dispersion (Average)
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15.75 to 16.25 KMC	SG 1516*	–6 DBM	SS 1516	5 mw	SA 1516	–70 DBM	45 MC
16.25 to 16.75 KMC	SG 1617*	–6 DBM	SS 1617	5 mw	SA 1617	–70 DBM	45 MC
18.0 to 22.0 KMC	SG 1822	–10 DBM	SS 1822	10 mw	SA 1822	–60 DBM	40 MC
22.0 to 25.0 KMC	SG 2225	–10 DBM	SS 2225	10 mw	SA 2225	–60 DBM	40 MC
24.7 to 27.5 KMC	SG 2427	–10 DBM	SS 2427	10 mw	SA 2427	–60 DBM	40 MC
27.27 to 30 0 KMC	SG 2730	–10 DBM	SS 2730	10 mw	SA 2730	–60 DBM	45 MC
29.7 to 33.52 KMC	SG 3033	–10 DBM	SS 3033	10 mw	SA 3033	–60 DBM	45 MC
33.52 to 36.25 KMC	SG 3336	–10 DBM	SS 3336	9 mw	SA 3336	–50 DBM	45 MC
35.1 to 39 7 KMC	SG 3540	–10 DBM	SS 3540	5 mw	SA 3540	–50 DBM	45 MC
37.1 to 42.6 KMC	External Source Power Measurement Range: +6 to +30 DBM Accuracy with Correction: ±2 DB		SS 3742	Approx. 3 mw			
41.7 to 50.0 KMC			SS 4150	Approx. 3 mw			

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Contributors

(Continued from page 104A)

W. W. Siekanowicz has been employed in the Tube Division of the Radio Corporation of America at Harrison, New Jersey since July of 1950 and has specialized in the development of traveling-wave tubes. His paper appears on page 1091 of this issue.

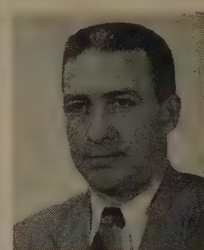
W. SIEKANOWICZ

He was born on January 3, 1928, in Poland. He received the B.S. degree in Electrical Engineering from the Imperial College of Science and Technology, London University, England in 1948, and the M.S. Degree in Electrical Engineering from Columbia University in 1950.

Mr. Siekanowicz is a member of Sigma Xi.

❖

N. W. Spencer (SM'50) whose paper appears on page 1104 of this issue is supervisor of engineering in upper atmosphere research at the ERI, University of Michigan.



N. W. SPENCER

He was born in Buffalo, N. Y. on August 4, 1918. He received the B.S. and M.S. degrees in electrical engineering from the University of Michigan.

From 1941 to 1944 he was employed as research engineer in resistance welding at Sciaky Bros., and from 1944 to 1946 he was a radar officer in the USNR. In 1946 he was at Cornell Aeronautical Laboratory for a short time. He left to become a research engineer in the ERI, University of Michigan. He assumed the position of project engineer in upper atmosphere research in 1947. He has also been engaged as a part time lecturer in E.E. laboratory since 1952.

Mr. Spencer is an associate member of Sigma Xi.

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For a photograph and biography of P. K. Tien, whose paper appears on page 1137 of this issue, see page 1676 of the November, 1953 issue of the PROCEEDINGS OF THE I.R.E.

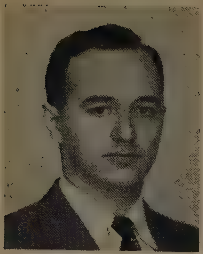
❖

H. M. Wachowski (S'51–A'53—M'53) in October, 1953, joined Armour Research Foundation of Illinois Institute of Technology where he has been engaged in radar research and development. He joined the staff of Northwestern University as an Assistant Professor of Electrical

(Continued on page 109A)

Contributors

(Continued from page 106A)



H. M. Wachowski

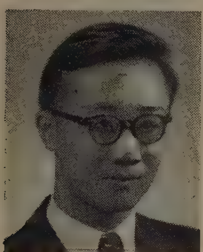
Engineering in 1952 teaching courses in field theory, microwaves, and electronics. His paper appears on page 1117 of this issue.

He was born in Chicago, Ill., on December 10, 1926. He received the B.S. and M.S. degrees in electrical engineering from Northwestern University in 1948 and 1950. From 1950 to 1952 he held an RCA Fellowship in Electronics at Northwestern University, where he received the Ph.D. degree in electrical engineering in 1952.

Dr. Wachowski is a member of Eta Kappa Nu and Sigma Xi.

❖

T. C. Wang whose paper appears on page 1117 of this issue entered Northwestern University in October 1952, and has been a research assistant working on the microwave space-charged vacuum tube detector project which is sponsored by the National Science Foundation, while at the same time he is studying for his Ph.D. degree.



T. C. Wang

He was born in China on December 22, 1927. He was a student at Chiao-Tung University in Shanghai before leaving China for this country. He received the B.S. and M.S. degrees in electrical engineering from State University of Iowa, Iowa City, Iowa in 1951 and 1952, respectively.

He is a member of Eta Kappa Nu and an associate member of Sigma Xi.

❖

G. W. Wood (A'51–SM'53) whose paper appears on page 1113 of this issue is at present Visiting Associate Professor of Physics at Tulane University, and a consultant to the National Research Council Committee on Undersea Warfare.

G. W. Wood

He was born in Warrensburg, Mo., on June 16, 1919. He received the B.S. degree in physics from Central Missouri State College in 1941. From 1943 until 1948 he was an instructor in the Department of Physics, Louisiana State University, receiving

(Continued on page 110A)

Contributors

(Continued from page 109A)

the M.S. degree in physics in 1946. In 1948 he entered the University of Texas for additional graduate study, and from 1948 to 1953 was a research physicist with the Defense Research Laboratory, University of Texas, engaged in underwater sound research, sponsored by the Navy Department.

Mr. Wood is a member of Sigma Xi, the Acoustical Society of America, and Sigma Pi Sigma.

❖

A. S. Zamanakos whose paper appears on page 1159 of this issue transferred in 1952 to the Naval Research Laboratory, where his present position is a member of the mathematics staff of Radio Division III. He was previously engaged in radio interference reduction work with the Bureau of Ships, Electronics Division, in Washington, D. C.

A. S. Zamanakos

He was born in Lowell, Mass., on February 25, 1924. He served in the U. S. Army from 1943 to 1946, being stationed in the United States and in Luzon, P.I. Mr. Zamanakos received his A.B. in mathematics from Boston University in 1950.

Section Meetings

Akron

Student paper competition: "The North American TRODI (Touchdown Rate of Descent Indicator)" by Donald Corbett; "The St. Lawrence Seaway," by James Wilson; "The Synchro Tie System," by Robert Savoy and "Electric Service to Ranch Type Homes," by Bruce Kent, all students, Akron University; March 9, 1954.

"Multichannel Magnetic Recorder for Physiological Functions," by J. F. Dobosy, Consulting Engineer, and Dr. W. L. Proudfit, Cleveland Clinic; April 22, 1954.

Atlanta

"Logical Design of Machines which Receive and Process Information for the Purpose of Automatic Control," by William Keister, Bell Telephone Labs.; April 13, 1954.

"Compatible Color Television and Its Relationship to the Broadcaster," by Dr. G. W. Brown, Director of Systems Research Lab., and Mr. C. N. Hoyler, RCA Labs.; May 14, 1954.

Baltimore

"What's to Drink?" by Dr. C. E. Renn, Johns Hopkins University; March 10, 1954.

"Unlimited Energy—Limited," by Dr. T. F. Nagey, The Glenn L. Martin Company; April 14, 1954.

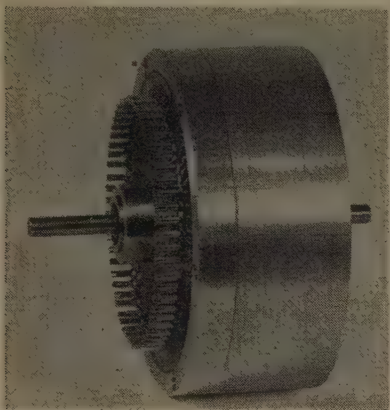
(Continued on page 112A)

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 80A)

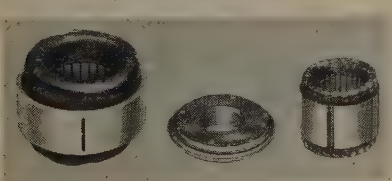
Multichannel Sampling Switch

General Devices, Inc., P.O. Box 253, Princeton, N. J., has a new high speed precision switch design which employs contact plates molded with high quality, mica-filled resin. It is relatively inexpensive to manufacture to customer specifications. Although the unit shown is only 2 inches in diameter and 1 inch long, the same technique applies to larger and to subminiature commutators of standard form used for telemetering, thermocouple and strain gauge sampling, parameter display, drift correction of multiple amplifiers, and so forth. The switch is available with any of a large variety of contact materials selected according to the application, and with a variety of features such as number of contacts, number of poles and sampling rates. It can be supplied with single or double-ended shaft or integrally mounted with its own driving motor.



Deflection Yokes for Radar

High precision radar deflection yokes are the newest products now being manufactured by **Constantine Engineerings Laboratories,** Mahwah, N. J.



Deflection yokes now in production include rotating and stationary types for PPI and rectangular displays. High-performance core materials such as Mu metal and Molly Permalloy are used. Specifications also include a wide range of inductances using complex winding distributions with high voltage insulations.

Specialized engineering data is available by writing the manufacturer. Give performance specifications in outline or drawing of type of yoke required.

(Continued on page 113A)

SECTION MEETINGS

(Continued from page 110A)

BEAUMONT-PORT ARTHUR

"Compatible Color Television and Its Relationship to the Broadcaster," by Dr. G. H. Brown, RCA; May 7, 1954.

BOSTON

"Coincident Current Magnetic-Core Memories," by W. N. Papian, Mass. Institute of Technology; February 18, 1954.

"Symphony Broadcasts on WGBH," by John Kessler, Mass. Institute of Technology Acoustics Lab., and Parker Wheatley, Hartford Gunn and Jordan Whitelaw of Lowell Institute Co-operative Broadcasting Council; March 18, 1954.

"Traveling Wave Tubes," by Dr. R. G. E. Hutter, Sylvania Physics Labs.; April 15, 1954.

BUFFALO-NIAGARA

Panel discussion: "Audio." Moderator: Prof. F. P. Fischer, University of Buffalo. Speakers: Ralph Black, Buffalo Philharmonic Orchestra; John Riggs, Rudolph Wurlitzer Co.; and F. H. Slaymaker, Stromberg-Carlson Company; May 19, 1954.

CEDAR RAPIDS

"Use of Dimensional Analysis in Heat Transfer Research," by Dr. Chia-Shum Yih, State University of Iowa; April 14, 1954.

"The Importance of Mechanical Design in Electronic Equipment," by W. R. Hewlett, Hewlett-Packard Co.; April 28, 1954.

CLEVELAND

"Multichannel Magnetic Recorder for Physiological Function," by J. F. Dobosy, Consulting Engineer; and Dr. W. L. Proudfit, Cleveland Clinic; April 22, 1954.

CONNECTICUT VALLEY

"The Electronic Organ," by G. H. Hadden and W. A. Johnson, Minshall-Estey Organ, Inc.; May 20, 1954.

DALLAS-FORT WORTH

"Telemetering—Instrumentation for Missile Testing," by Ira Heimlich, Chance Vought Aircraft; "Under Water Sound Measurement on Dunking Sonar," by W. D. Penn, Texas Instruments; April 29, 1954.

DENVER

"The N.T.S.C. Color Television System and It's Background," by W. R. Hewlett, President, IRE; April 2, 1954.

"The National Bureau of Standards Atmospheric Radio Noise Program," by W. Q. Critchlow, National Bureau of Standards; April 23, 1954.

EL PASO

"Management of the Glen L. Martin Co. and the Viking Missile," by W. G. Purcy, Project Manager on the Viking Missile at White Sands Prov. Grds.; April 28, 1954.

EMPORIUM

"Philosophy of Transistor Circuit Design," by R. F. Shea, General Electric Company; April 20, 1954.

"Penn State Nuclear Reactor," by Dr. W. M. Breazeale," Penn. State; May 18, 1954.

EVANSVILLE-OWENSBORO

"Color Television," by Dr. W. R. G. Baker, Vice President, General Electric Company; May 18, 1954.

FORT WAYNE

"Automatic Blind Landing Guidance Systems," by D. R. Treffeison, Sperry Gyroscope Company; May 6, 1954.

(Continued on page 114A)

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(*Continued from page 111A*)

Printed Circuit Bulletin

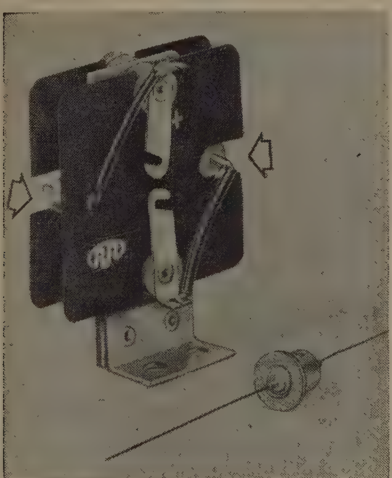
Insulated Circuits, Inc., 115 Roosevelt Ave., Belleville, N. J., has announced the availability of their new color bulletin No. 106, on Carry-Through Printed Circuits.

The new bulletin discusses many facets of printed circuit production, including insulating base materials with respect to dip-solderability, insulation resistance under operating conditions, thermal shock coefficients, tensile strength, dielectric strength, chemical resistance and fabrication limits. Conductive patterns regarding the choice of metal, thickness of the pattern, line definition and current capacity in amperes, and plating of conductive patterns depending on the applications of the printed circuit, either "Printed Wiring" or "Printed Switching Patterns," are also discussed.

The bulletin includes detailed photographs and drawings on various printed circuits and patterns. Also listed are I.C.I.'s services. For a copy of the bulletin write to the company.

Junction Power Diodes

Seven new junction power diodes have just been added to their line by **Radio Receptor Co., Seletron & Germanium Div.,** 251 W. 19th St., New York 11, N. Y. No. 1N91, and 1N151, are diodes as shown in the foreground of the illustration. No. 1N151, 1N152, and 1N153 are single diodes mounted on a cooling fin, while No. 1N158, illustrated, consists of two diodes as indicated by arrows, connected in series and mounted on a cooling fin.



These units are quite small, and are hermetically sealed. The Type 1N153 has a peak inverse voltage rating of 300 and is capable of delivering 0.5 ampere into a resistive load with a voltage drop of only 0.7 volt.

For further information about these junction power diodes, including complete characteristics and curves, communicate with the Sales Dept. of Radio Receptor.

(*Continued on page 141A*)

SECTION MEETINGS

(Continued from page 112A)

Hamilton

"Engineering Management in the Electronics Industry," by H. E. Rice, Canadian Westinghouse Company; May 10, 1954.

Houston

"A Fourier Analyzer by Cross-Correlation Methods," by D. K. Davis, N. E. Hermann, Students, University of Houston; "The Flux Gate Magnetometer," by James Nygaard, Student, Texas A & M; April 20, 1954.

Huntsville

"Color Television," by L. E. Rawls, WSM-TV; April 30, 1954.

Inyokern

"Servomechanisms," by F. L. Moseley, F. L. Moseley Company; April 12, 1954.

Tapescript: "How Much Distortion Can You Hear?" and talk, "Data Handling System at the Air Force Missile Test Center, Fla.," by R. B. Bonney, Electronic Engineering Company of Calif.; May 10, 1954.

Ithaca

"Color Television," by I. C. Abrahams, General Electric Color Television Unit; April 23, 1954.

Little Rock

"Measurements in Communications," by N. B. Fowler, American Tel. and Tel. Co.; film, "Application of 'Tinkertoy' Techniques to Manufacture of Electronic Equipment"; May 11, 1954.

Long Island

"NTSC Color Television Receiver Design," by I. E. Lempert, Westinghouse Electric Corp.; April 21, 1954.

London

"The Technical Aspects of a Modern T.V.," by G. Robitaille, Section Chairman, and tour of Radio Station CFPL-TV; April 24, 1954.

Los Angeles

"Predictors: Design of Systems from the Predictor and Energy-storage Viewpoint," by O. J. M. Smith, faculty, University of California; and "The Engineering Personality," by Floyd Graham, North American Aviation, Inc. Dinner speaker: "A Program of Self Post-graduate Training," by Dr. Simon Ramo, The Ramo-Woolridge Corp.; April 6, 1954.

"Information Theory—What Is It and Where Is It Going?" by W. G. Tuller, Melpar, Inc.; and "A New Mt. Wilson Television Transmitting Antenna Installation," by C. G. Pierce, American Broadcasting Company, and J. A. Stagnaro, Station KABC-TV, ABC; and film, "Wanna Buy A Record?" May 4, 1954.

Louisville

Plant tour of Southern Bell Tel. and Tel. Company under the direction of R. S. Watson, District Manager, and S. H. Gates, Div. Transmission Engineer; May 13, 1954.

Miami

"Manufacturing Techniques for Quartz Crystals," by R. E. Bassett, Jr., Rex Bassett, Inc.; April 2, 1954.

"Broad Band Transmission," by W. H. Doherty, Bell Telephone Labs.; May 7, 1954.

(Continued on page 117A)

SECTION MEETINGS

(Continued from page 114A)

Montreal

"Flight Simulator," by Dr. H. F. Quinn, Research Director, IRE; April 21, 1954.

Symposium on Local Engineering Developments: "Oscillo-Meter-Scope No. 104," by G. K. Davidson, Canadian Aviation Electronics Ltd.; "Portable Insulation Tester and Earth Resistance Tester," by H. H. Schwartz, Electrodesign; and "Electronic Oximeter," by R. H. Taplin, Canadian Marconi Company; May 19, 1954.

New Orleans

"Operations in a Shipboard Combat Information Center," by Cmdr. W. R. Sherman, USN; April 27, 1954.

New York

"Characteristics and Applications of Microstrip Lines for Microwave Wiring," by Maurice Arditi, Federal Telecommunication Labs.; April 7, 1954.

North Carolina-Va.

"Printed Circuits," by Robert Lumley, student, North Carolina State College; "The NTSC Color Television System," by Henry Skutt, student, Virginia Polytechnic Institute; "A Pulse Amplitude Discriminator," by Ted Haggi, student, North Carolina State College; and "High Voltage Test Equipment," by T. E. Bowman, student, University of Virginia; May 7, 1954.

Ottawa

"Microwave Scanners," by Dr. Harry Gruenberg, National Research Council; April 22, 1954.

Philadelphia

Westinghouse demonstration: "Energy in Action" and talk, "Radio Signals from the Milky Way," by B. J. Bok, Agassiz Station Project in Radio Astronomy; May 6, 1954.

Phoenix

"Electrical Properties of Capacitors and Their Significance," by W. S. Franklin, John E. Fast and Company; February 9, 1954.

"Engineering Developments in the West," by Beardsley Graham, Stanford Research Institute; and "Microwave Measurements," by B. P. Hand, Hewlett-Packard Co.; February 26, 1954.

Tapescript: "Physics of Music and Hearing," by W. E. Koch, Bell Telephone Labs.; March 5, 1954.

"The Physical and Biological Effects of Air Ionization," by Dr. T. L. Martin, Jr., University of Arizona; March 26, 1954.

Pittsburgh

"Ionic Oscillators," by David Gross, student, West Virginia University; "A High Intensity Light Source for Ultra-High Speed Photography," by Philip Eckman, student, Carnegie Inst. of Technology; and "Television Interference," by G. R. Williams, student, West Virginia University; April 12, 1954.

Portland

"The Method of Vibration Pickup in Wurlitzer Organs," by Edwin Baer, Wurlitzer Company; April 22, 1954.

Princeton

"Transistor Circuits and Applications," by G. C. Sziklai, RCA Labs.; April 22, 1954.

Rome-Utica

"Electronic Aids for the Deaf," by Fred Sparks, N. Y. School for the Deaf; and "Important Design Considerations for Multiple Broadcast Television Antenna Installations," by R. H. Wright, RCA; May 4, 1954.

(Continued on page 118A)

SECTION MEETINGS

(Continued from page 117A)

Sacramento

Election of officers; May 14, 1954.

St. Louis

Panel discussion: "Improvement of Airborne Electronic Equipment." Moderator: Cdr. C. H. S. Murphy, USN. Speakers: A. R. Anderson, McDonnell Aircraft Corp.; J. L. Wimpey, McDonnell Aircraft Corp.; E. W. Logan and F. F. Rohne, Emerson Electric; March 4, 1954.

Student competition: "Electronic Voltage Regulator," by T. Bryan and R. Lauer, Washington University; "Vacuum Tube Characteristic Plotter," by R. C. Distler, St. Louis University; "Re-entrant Flux," by B. Douglass, Washington University; "An Automatic Antenna Control," by K. Walton, University of Mo.; "Stereophonic and Binaural Sound," by M. Barylski, Mo. School of Mines and Metallurgy; "Discussion of Magnetic Amplifiers" by C. Johnson, Mo. School of Mines and Metallurgy; and "Automatic Test of DC Generators," by J. Cooney, St. Louis University; April 29, 1954.

San Antonio

"Instrumentation and Control of Physical Systems," by Dr. J. D. Trimmer, University of Tennessee; April 14, 1954.

"Color Television," by Dr. G. H. Brown, RCA Labs.; May 4, 1954.

San Diego

"Decade Counter Tubes," by Kenneth Buzzard, student, San Diego State College, and "High Voltage Demonstration Apparatus," by Lambert Dolphin, student, San Diego State College; May 4, 1954.

Schenectady

"Color Television," by Paul Howells, General Electric Company; April 21, 1954.

Tulsa

"Instrumentation for Electrical and Visual Observation of a Three Dimensional Seismic Wave Model," by J. W. Miller and "Role of the Card Programmed Electrical Calculator in Scientific Computations," by W. B. Rider, both engineers, Standoline Research Center; April 22, 1954.

Twin Cities

"Research Products of the DuPont Company," by P. R. Leach, DuPont Company; May 5, 1954.

Vancouver

Tapescripts: "A Single Ended Push Pull Amplifier," by Dr. Peterson and Dr. D. B. Sinclair, General Radio Company; and "The Sound Survey Meter," by Dr. Peterson; April 26, 1954.

Washington, D. C.

"UHF TV Broadcast System—Problems and Prospects," by G. E. Sterling, F.C.C.; W. J. Morlock, General Electric; and Dr. Wen Yuan Pan, RCA; May 10, 1954.

Williamsport

"The Present Status of the Capacitor and Resistor Art with Respect to Electronic Components," by Leon Podolsky, Sprague Electric Company; February 17, 1954.

"Color Television," by Charles Brough, Westinghouse Corp.; March 31, 1954.

SUBSECTIONS

Berkshire County

"Punched Card Operated Airborne Navigation Computer," by E. B. Thornley, Collins Radio Company; April 26, 1954.

(Continued on page 120A)

Section Meetings

(Continued from page 118A)

MID-HUDSON

"Specialized Ferrite Applications," by Ephrain Gelbard, General Ceramics and Steatite Corp.; March 18, 1954.

"Development Engineering Management," by W. C. Tinus, Bell Telephone Labs.; April 22, 1954

MONMOUTH

Field trip to Long Lines Dept., American Tel. and Tel. Co.; March 13, 1954.

NORTHERN NEW JERSEY

"Advances in Semiconductor Power Rectifiers," by A. C. Sheckler, General Electric Company; March 10, 1954.

"Miniature Lacquer Film Capacitors," by D. A. McLean and H. G. Wehe, Bell Telephone Labs.; April 14, 1954.

ORANGE BELT

"A Magnetic Video Tape Recorder," by J. T. Mullin, Crosby Enterprises; "Physics of Sound and Hearing" (Tapescript) by W. E. Kock, Bell Telephone Labs.; April 14, 1954.

PALO ALTO

"Engineering Aspects of the Tinkertoy Process," by R. L. Miller and R. E. Bauer, USN Postgraduate School; April 16, 1954.

"Transistor Oscillators," by Dr. Hans Hollman, U. S. Naval Air Missile Test Center, Point Mugu; March 3, 1954.

TUCSON

"IRE Functions and Student Graduate Electronic Training," by Dr. Pettit, Regional Director of IRE Region 7, University of Stanford; April 29, 1954.

USAFIT

"Transmission Lines, Wave Guides and Connectors," by G. B. Eley, WADC; April 30, 1954.

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. .... The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
1 East 79th St., New York 21, N.Y.

ENGINEERS

Manufacturer of electrical indicating instruments requires men with experience in design of D'Arsonval Dynamometer and Iron Vane type meters. Engineering degree not essential. Experience will qualify. In reply give full details of education and experience and salary requirements. Box 768.

SPECIAL PROJECTS ENGINEER

Our newly established Field Service and Technical Training Unit is preparing for expanding operations. We need an electronics engineer or physicist who enjoys a challenging variety of work with advanced types of electro-mechanical sys-

(Continued on page 122A)

MEMBERSHIP

The following transfers and admissions were approved to be effective as of June 1, 1954:

Transfer to Senior Member

Anner, G. E., 103 S. Busey Ave., Urbana, Ill.
Anthony, R. L., 576 W. Englewood Ave., West Englewood, N. J.
Bain, J. D., Canadian Arsenals, Ltd., Leaside, Ont., Canada
Barker, J., 142 Farmington Rd., Utica 3, N. Y.
Bedrosian, S. D., 993 Ridge Ave., Manasquan, N. J.
Bennett, A. F., Countryside, Summit, N. J.
Braverman, N., 214 Lexington Ave., Dayton 7, Ohio
Brewer, G. R., 3704 Via Cardelina, Palos Verdes Estates, Calif.
Brewer, W. L., 275 Sagamore Dr., Rochester 17, N. Y.
Brockway, W. W., 2308 Beloit Ave., Los Angeles 64, Calif.
Coffman, B. C., 20 Chester Ave., Massapequa, L. I., N. Y.
Cone, D. R., 6017 Chabolyn Ter., Oakland 18, Calif.
DeVore, C., 161 Danbury St., S.W., Washington 24, D. C.
Dunnigan, F. A., 5109 Chesley Ave., Los Angeles 43, Calif.
Dutton, O. B., 513 Calle de Arboles, Redondo Beach, Calif.
Elliott, G., 3556 Grey Ave., N.D.G., Montreal, Que., Canada
Field, P. A., 369 Pleasant Park Rd., Ottawa 1, Ont., Canada
Foley, W. V., 12 Left Wing Dr., Baltimore 20, Md.
Frank, R. L., 23 Red Brook Rd., Great Neck, L. I., N. Y.
Gere, A. J., 126 Edgewood Rd., Westwood, Mass.
Giles, J. W., 37 Cambridge Rd., R.F.D. 1, Whitesboro, N. Y.
Goldfus. S. M., S. M., 7723 S. Essex, Chicago 49, Ill.
Gruber, J. R., 1397 Congress St., S.E., Washington 20, D. C.
Hadley, A. M., 73 Highland Ave., Chatham, N. J.
Honer, R. E., 5462 Mary Lane Dr., San Diego 15, Calif.
Jasik, C., 44 Overlook Rd., Great Neck, L. I., N. Y.
Kirkpatrick, W. E., Bell Telephone Laboratories, Inc., Murray Hill, N. J.
Klepinger, L. H., 11417 E. Balfour St., Whittier, Calif.
Lang, H. M., 2606 E. 33 Pl., Tulsa 5, Okla.
Lantz, P. A., 521 Oakwood St., S.E., Washington 20, D. C.
Lotz, W. E., Jr., 1324 N. Illinois St., Arlington 5, Va.
Maher, R. A., 6133 Sunridge Dr., Cincinnati 24, Ohio
Massell, E. M., Electronic Associates, Inc., Long Branch, N. J.
McFarlane, D. J., 2 Bonad Rd., Winchester, Mass.
McMillen, R. C., 200 W. Pembrey Dr., Wilmington, Del.
Meador, B. M., 604 W. 86 Ter., Kansas City 14, Mo.
Moreno, C. A., 77 Dosoris La., Glen Cove, L. I., N. Y.
Neidert, J. H., 9 Surrey Rd., New Hyde Park, L. I., N. Y.
Nichols, B., School of Electrical Engineering, Cornell University, Ithaca, N. Y.
Nikonenko, P. V., 1730 Pacheco St., San Francisco 16, Calif.
Nishino, H. H., 531 Kings Highway, Apt. R-4, Moorestown, N. J.
Odell, N. H., Bell Telephone Laboratories, Inc., 555 Union Blvd., Allentown, Pa.

(Continued on page 136A)

MEMBERSHIP

(Continued from page 135A)

O'Neill, L. H., 632 W. 125 St., New York, N. Y.
Peterson, R. A., 7245 Roe Ave., Prairie Village 15, Kans.
Pipes, L. A., University of California, Los Angeles 24, Calif.
Radcliffe, F. E., 205 Washington Ave., Chatham, N. J.
Roney, E. L., 159 Orange Dr., San Luis Obispo, Calif.
Samuelson, C. D., 271 Midland Ave., River Edge, N. J.
Schwarz, R. J., Electrical Engineering Department, Columbia University, New York, 27, N. Y.
Smith, J. S., 1995 Sedgwick Ave., New York 53, N. Y.
Smith, P. G., 54 Daffodil La., Wantagh, L. I., N. Y.
Snyder, W. W., 954 N. Forest Rd., Williamsville 21, N. Y.
Swarm, H. M., 6856–19, N.E., Seattle 5, Wash.
Tims, E. F., Electrical Engineering Department, Washington University, St Louis 5, Mo.
Torgow, E. N., 80 Knolls Crescent, New York 63, N. Y.
Warner, A. H., 3321 Grayburn Rd., Pasadena 10, Calif.
Wildes, K. L., Massachusetts Institute of Technology, 77 Massachusetts Ave., Cambridge 39, Mass.

Admission to Senior Member

Barlow, H. M., 12 Higher Dr., Banstead, Surrey, England
Barnes, T. G., Texas Western College, El Paso, Tex.
Bird, J. R., Russell Rd., Chagrin Falls, Ohio
Boyd, H. R., 44 Lincoln St., Stoneham, Mass.
Brite, L. A., Box 285, R.F.D. 2, Atwater, Ohio
Crone, R. L., 7909 Alverstone Ave., Apt. 2, Los Angeles 45, Calif.
Court, P. R. J., 393 Eggert Rd., Buffalo 15, N. Y.
Dench, E. C., 795 Charles River St., Needham 92, Mass.
Drvostep, J. J., 424 E. 75 St., New York 21, N. Y.
Gibson, V. R., Jr., 120 Lynn Dr., North Syracuse, N. Y.
Hamilton, C. R., 38 Le Britton St., Locust Valley, L. I., N. Y.
Hinsdale, J. G., 921 Ballard, S.E., Grand Rapids, Mich.
Hogan, C. L., 18 Battle Green Rd., Lexington 73, Mass.
Hull, H. L., Capehart-Farnsworth Co., 3702 E. Pontiac St., Fort Wayne, Ind.
Jaynes, E. T., Microwave Laboratory, Stanford, Calif.
Jones, R. W., 2618 Orrington St., Evanston, Ill.
Kundel, T., Frandford Arsenal, Quarters 6, Philadelphia 37, Pa.
Levine, B., Ketay Manufacturing Corp., 555 Broadway, New York 12, N. Y.
Levine, I., 4201 Massachusetts Ave., N.W., Apt. 281, Washington 16, D. C.
Merrill, J. R., 27 York St., Lexington, Mass.
Miller, M. A., 344 Prospect Ave., Little Silver, N. J.
Neprud, R. N., 14 Roosevelt Ave., Lancaster, N. Y.
Nunan, C. S., 1460 Summit Rd., Berkeley 8, Calif.
Oberle, G. W., 8862 Belair Rd., Baltimore 14, Md.
Patterson, T. C., 2638 Harrison St., Arlington, Calif.
Powell, R. C., 140 N.E. 95 St., Miami Shores, Fla.
Pulvari, C. F., 2014 Taylor, N.E., Washington 18, D. C.
Rabinow, J., Rabinow Engineering Co., 7212 New Hampshire Ave., Takoma Park 12, Md.
Reid, C. R., 2432–16 St., Cuyahoga Falls, Ohio
Roggenbuck, R. A., 25400 Ingleside Dr., R.F.D. 3, Birmingham, Mich.

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MEMBERSHIP

(Continued from page 136A)

Shapp, M. J., 626 S. Bowman Ave., Merion Station, Pa.
Torre, A. J., 139 N. Childs St., Woodbury, N. J.
Witty, W. M., 4228 Lovers La., Dallas 25, Tex.
Young, F. M., Acoustics Laboratory, 20C-006, Massachusetts Institute of Technology, Cambridge 39, Mass.

Transfer to Member

Adler, L. L., 69-22-225 St., Bayside 64, L. I., N. Y.
Arneth, A. P., Jr., I.B.M. Corp., Plant 2, Department 539, Poughkeepsie, N. Y.
Bernstein, G., 437 Miller Ave., Brooklyn 7, N. Y.
Berntsen, P. L., 126 Fifth Pl., Manhattan Beach, Calif.
Busser, J. H., 6320 Ross St., Philadelphia 44, Pa.
Drosin, A. B., 418 St. Johns Pl, Brooklyn 38, N. Y.
Ellis, T. E., Jr., R.F.D. 4, Rome, N. Y.
Estin, A. J., 2631 Ninth St., Boulder, Colo.
Felsman, W. O., Jr., 2711 Robindale Ave., Akron 12, Ohio
Finnegan, F. J., 115 Bright St., Waltham 54, Mass.
Frank, M. E., 1607 Flower St., Glendale 1, Calif.
Frost, H. B., 46 Westgate, Cambridge 39, Mass.
Garrigus, W. E., 26 N. Edmondson Ave., Indianapolis 19, Ind.
Griffith, B. D., 461 N. Forklanding Rd., Maple Shade, N. J.
Hand, G. C., Jr., 1603 E. Gowen Ave., Philadelphia 19, Pa.
Jacobson, R., 15243 Gager St., Pacoima, Calif.
Joyce, K. J., 159-20-65 Ave., Flushing, L. I., N. Y.
Kamat, V. Y., 12, Dwarkadas Mansion, Sandhurst Rd., Bombay 4, India
Kannan, S., E. K. Cole, Ltd., Development and Engineering Department, Southend-on-Sea, Essex, England
Kilpatrick, D. G., 3527 Indian Queen La., Philadelphia 29, Pa.
Koons, E. E., 30 Wynnwood Ave., Tonawanda, N. Y.
Lambert, T. A., 201 Thor Dr., Falls Church, Va.
Lavin, H. D., 517 Wayside Rd., Neptune, N. J.
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Oda, D. J., 213 Johnson Ave., Runnemede, N. J.
Read, A. A., 511 Northwestern Ave., Ames, Iowa
Ross, J. M., Box 2371, Stanford, Calif.
Schaffer, L. I., 1207 Gracewood Dr., Greensboro, N. C.
Seymour, H. A., 9715 Sutherland Rd., Silver Spring, Md.
Shaw, N. H., 92 W. Calaveras St., Altadena, Calif.
Spencer, J. C., 1524 Shirley Dr., New Orleans 14, La.
Spirer, H. F., 122 Retford Dr., Liverpool, N. Y.
Spiro, N. J., 9 MacDonald St., Clifton, N. J.
Test, L. D., 1509 Birch St., Richland, Wash.
Thompson, C. W. N., 4565 Lesher Dr., Dayton, Ohio
Trapp, J. A., 3238 Josie Ave., Long Beach 8, Calif.
Westcoat, W. L., 110 Oakwood Dr., Packanack Lake, N. J.

Admission to Member

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Artman, R. L., 408 West South, Angola, Ind.
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Barry, H. P., Jr., 89-46 Gold Rd., Ozone Park, L. I., N. Y.
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Beeh, R. C. M., 247 E. 51 St., New York 22, N. Y.

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MEMBERSHIP

(Continued from page 138A)

Bennett, R. E., 2109 Colfax, Evanston, Ill.
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Chamberlain, W. E., Jr., 16 Elliott Ave., New London, Conn.
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Granberry, H. W., 5566 Aspen St., Bellaire, Tex.
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Johnson, R. E., 19 Governor La., Schenectady, N. Y.
Johnson, R. F., Adalia, Ltd., 20 Carleton St., Toronto, Ont., Canada
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Leunis, J., 1850 General Urquiza, Florida, FCNGBM, Buenos Aires, Argentina
Lowenberg, E. C., 521 Holt Ave., Iowa City, Iowa
Luce, R. R., 107 Maurine Dr., San Antonio 10, Tex.
Macdonald, B., 11 Woodside Rd., Sutton, Surrey, England
McCollough, E. C., RCA Field Engineer, 181 Sig Dep Co., APO 971, c/o Postmaster, San Francisco, Calif.
McConnell, H. M., 355 Idlewood Rd., Pennridge, Pittsburgh 35, Pa.
Menke, W. W., 185 Maple Ave., Rockville Centre, L. I., N. Y.
Michel, D. H., 1496 Sauve (East), Apt. 1, Ahuntsic, Montreal, Que., Canada
Millar, H. D., 1200 M. D. Anderson Blvd., Houston 5, Tex.
Molinari, H. A., 11 Leonhard-Ragazweg, Zurich, Switzerland
Murphy, P. J., 201 State Life Bldg., New York Central System, Indianapolis, Ind.
Nearing, F. G., Box 365, R.F.D. 1, Kimberly Rd., Barrington, Ill.
Northover, F. H., 12 Kent Pl., St. John's, Newfoundland, Canada

(Continued on page 140A)

MEMBERSHIP

(Continued from page 139A)

Ogden, D. R., 1258 Fitzgerald Ave., San Francisco 24, Calif.
Olsen, H. G., 9300 Highway 54, El Paso, Tex.
Osder, S. S., 33 Deer La., Wantagh, L. I., N. Y.
O'Shaughnessy, C. A., 44th A & E Sqdn., Lake Charles AFB, Lake Charles, La.
Pastoriza, H. G., Jr., 10 Oriole Ave., Bronxville 8, N. Y.
Phaneuf, C. A., 54 Wyman St., Waban 68, Mass.
Polen, E., 859 Avenue Z, Brooklyn 23, N. Y.
Pridgen, W. L., 1424 N. Rolling Rd., Baltimore 28, Md.
Pulido, G. G., Jr., 803 Floyd Ave., Rome, N. Y.
Quigley, G. T., 7 Kingswell Crescent, Scarborough, Toronto 16, Ont., Canada
Rawls, J. W., Jr., 7348 Franklin Rd., El Paso, Tex.
Rester, G. F., 455 Peters Blvd., Brightwaters, N. Y.
Roberto, S. N., 50-39–184 St., Flushing 65, L. I., N. Y.
Rosenbaum, B., Bldg. 13-3, RCA Victor Division, Camden, N. J.
Rulong, C. P., R.F.D. 2, Moundsville, W. Va.
Rush, S., 73-63 Bell Blvd., Bayside 64, L. I., N. Y.
Ryskamp, J. H., 19340 Frazier Dr., Rocky River 16, Ohio
Siering, R. K., 3219 W. 25 St., Cleveland 9, Ohio
Sieron, E. M., 1129 Elbank Ave., Baltimore 12, Md.
Smolen, A., 15-11 Chandler Dr., Apt. 21, Fair Lawn, N. J.
Spaulding, G. B., 10434 Sylvia Dr., Dallas, Tex.
Steel, D. L., Sr., 710–14 St., N.W., Washington 5, D. C.
Stubner, F. W., 555 Union Blvd., Allentown, Pa.
Styren, K., Telefonaktiebolaget, L. M. Ericsson, Stockholm 32, Sweden
Suita, T., Rm. 4-242, Massachusetts Institute of Technology, Cambridge 39, Mass.
Sullivan, J. F., 72 Hawley St., Lawrence, Mass.
Tavukcu, G. D., 52 W. 93 St., Apt. 2E, New York 25, N. Y.
Timmers, A. G. W., 45 Elizabeth St., Burlington, Ont., Canada
Unger, J. H., 4930 Idaho St., Houston 21, Tex.
Walters, L. R., I.B.M. Corp., Plant II, South Rd., Poughkeepsie, N. Y.
Watt, A. D., National Bureau of Standards, Boulder, Colo.
Weigert, H. A., 375 Riverside Dr., New York 25, N. Y.
Youngquist, R. J., 1230 Rose Vista Ct., St. Paul 13, Minn.

The following elections to the Associate grade were approved to be effective as of June 1, 1954:

Aghib, E. G., 42 Charlton St., New York 13, N. Y.
Agnitch, J., c/o Emerson Electric Manufacturing, 410 W. First St., Dayton 2, Ohio
Ahern, H. G., 3138 S.W. 23 St., Miami 34, Fla.
Albert, B., 98 Raymond St., Cambridge 40, Mass.
Altman, J. L., 154 Chapman St., Watertown 72, Mass.
Andrews, S. S., Jr., 1325 State St., Eau Claire, Wis.
Anthony, J. C., 330 W. 42 St., New York 36, N. Y.
Appelman, J. F., 46 Meehan Dr., Page Manor, Dayton 3, Ohio
Archer, G. R., 46 Ford St., Lynn, Mass.
Arnold, W. R., 5117 Old Frederick Rd., Apt. E5, Baltimore 29, Md.
Auclair, M., 402 Egan, Verdun, Que., Canada
Balderson, W. M., Jr., Department of Army, Connecticut Ave., Van Ness St., N.W., Diamond Ordnance Fuze Laboratory Bldg., Washington, D. C.
Balk, D. B., 59 Grant St., Utica, N. Y.

(Continued on page 142A)

PROFESSIONAL ENGINEERING CARDS

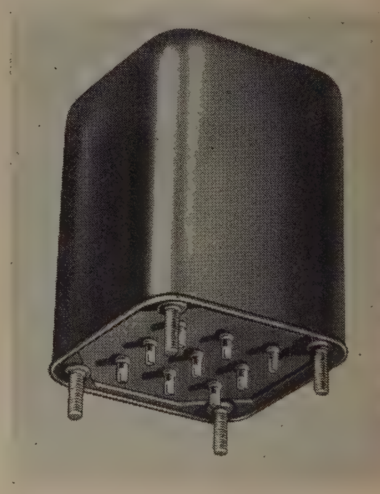
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 113A)

High-Fidelity Output Transformer

A new high-fidelity output transformer, BO-13, designed for ultra-linear operation, has been announced by the **Chicago Div. of Chicago Standard Transformer Corp.**, 3501 Addison St., Chicago 18, Ill.



This new transformer features Chicago's "sealed-in-steel" construction. The drawn steel case measures $3\frac{11}{16} \times 3\frac{5}{16} \times 4\frac{11}{16}$ inches. It has pin-type terminals and is designed with studs for flush chassis mounting.

Complete details, including performance curves, are given in Chicago Bulletin #33, available free from any Chicago distributor or by writing the company direct

MEMBERSHIP

(Continued from page 140A)

Barker, W. M., General Electric Co., Electronics Park, Syracuse. N. Y.
Bauer, S. J., Box 571, R.F.D. 1, Belmar, N. J.
Baxley, W. E., Box 152, R.F.D. 3, Texarkana, Tex.
Baxter, S., Radio Range Station, Beatton River Airport, Via Fort St. John, B.C., Canada
Beatty, D. L., 1616 Westport Rd., Kansas City 11, Mo.
Berta, E. J., 21305 Frazho, St. Clair Shores, Mich.
Bhattacharyya, B. K., P-34 A. Matijheel Ave., Dum Dum, Calcutta 28, West Bengal, India
Bickwoldt, W. H., 10733 Phinney Ave., Seattle 33, Wash.
Bissonette, A. G., 228 Niagara Ave., Dayton 5, Ohio
Blake, G. C., 4711 Eighth St., N.W., Washington, D. C.
Booth, W. C., 129 Eighth St., Federal Electric Co., Passaic, N. J.
Bowker, I. H., 227 Mann Ave., Fairborn, Ohio
Brathwaite, A., 300 Sutter Ave., Brooklyn 12, N. Y.
Brodigan, H. F., 245 W. Fourth St., New York 14, N. Y.
Bronstein, L. M., 64-35 Yellowstone Blvd., Forest Hills 75, L. I., N. Y.
Brunhart, W., c/o P. Wolfe, 61 Carroll St., Springdale, Conn.
Burneka, C. R., 6335 N. Dixie Dr., Dayton 4, Ohio
Butler, J. W., 13 Riverside Pl., Scotia 2, N. Y.
Cairns, H. E., Church St., Bernardston, Mass.
Canaday, H. T., 148 Lexington Ave., Dayton 7, Ohio
Carlotto, A. J., 57 Orlando St., West Haven 16, Conn.
Czrpenter, R. L., 3146 Rowena Ave., Los Angeles 27, Calif.
Carnahan, K. R., 349 Wayne Dr., Fairborn, Ohio
Catanzarite, F. J., 5650 Gross Dr., Dayton 3, Ohio
Christgau, M. A., 11816 Lakewood Blvd., Downey, Calif.
Clark, J., 298 Empire Ave., St. John's, Newfoundland, Canada.
Coolidge, C. A., Jr., 94 Somerset St., Belmont 78 Mass.
Cote, L. J. A., 2060–18 St., Quebec City, Que., Canada
Cox, R. E., 1361 Lecaron, Sarnia, Ont., Canada
Curley, J. C., 32 Ivy Home Rd., Hampton, Va.
Cushing, W. J., 434 Grand Ave., Apt. 22B, Dayton 5, Ohio
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